Chapter-1

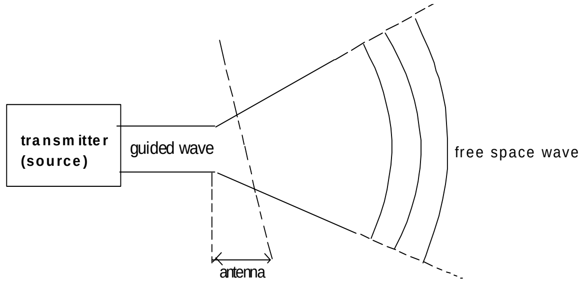


fig.(i) antenna as a transtion device

Antenna:- Antenna, a aerial is a metallic elevated conductor (as a rod or wire) for radiating or receiving g radio waves. It is a structure between free space and a guiding device as shown in fig[1]. The guiding device or transmission line may take the form, of co-axial line or a hollow pile (wave guide) and is used to transport electromagnetic energy from the transmitting source to the antenna or from receiving antenna to the receiver.

Types of antenna:-

Following are the major types of antenna.

(i) Wire antenna:-

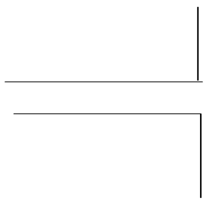


fig. dipole

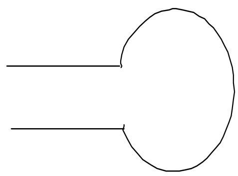


fig. circular loop

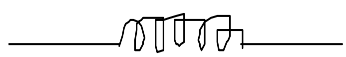


fig. Helix

Wire antennas are familiar to everyone because they are seen and used everywhere in buildings, automobiles, ships, spacecrafts, aircrafts and so on. There are various shapes of wire antenna such as straight wire (dipole) circular loop wire, Helix and so on.

(ii) **Aperture antenna**:- An antenna having an aperture (opening) with a certain geometrical shape is referred as an aperture antenna . it is in use because of utilization of higher frequencies. The aperture may take the form of wave guide or a Horn. Antenna of this type are very useful for aircraft and sac raft and space craft application because they can be very conveniently mounted on the skin of the aircraft or spacecraft. In addition they can be covered with dielectric material to protect them from. Hazar dous condition of the environment.

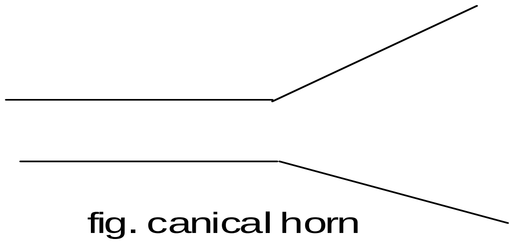
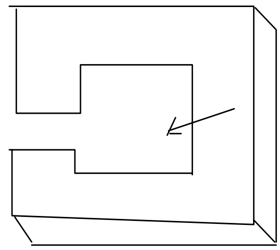


fig. canical horn

fig.waveguide

(iii) **Microstrip antenna:-**



ground plane

fig. raectangular microstrip antenna

An antenna having metallic patch on a grounded substract as shown in fig known as microstrip antenna. The patch may take the form of circular or rectangular. There are used because of their attractive. Radiation characteristics specially low-confor table for planar and non planar surface, simple and in expensive to fabricate. These antennas can be mounted on the surface of high performance aircraft, spacecraft, satellite, missiles and even in hand held mobile.

(iv) **Array antenna**:- An array antenna is an assembly of radiating elements in an electrical and geometrical configuration. Array antennas are used because many applications require radiation characteristics that may not achievable by a single element. The term array is reserved (used) for an arrangement in which individual radiators are separated as shown in fig.

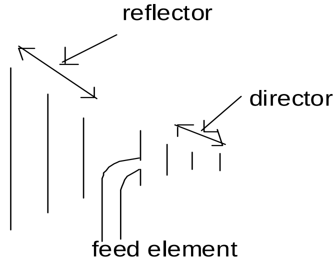


fig. simple array antenna

(v) **Reflector antenna:-**

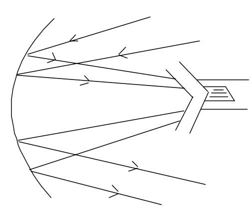
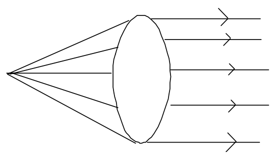


fig. parabolic antenna.

Reflector antenna take many geometrical configurations. Some of the most popular configuration are the planar, corner and parabolic. Because of the need to communicate over great distances, sophisticated forms of antennas had to be used in order to transmit and receive signals that had to travel millions of miles. This type of antenna may have the diameter as larger 306m.

(vi) Lens antenna:-



Lens antennas are used to collimate incident divergent energy to prevent it from spreading in undesired direction by properly shaping the geometrical configuration and choosing the appropriate material of the lens antennas are used in most of the same applications as the parabolic reflectors specially at higher frequencies.

Review of electromagnetic waves and equations:-

Maxwell's equations for steady electromagnetic field:-

Differential form	**Integral form**
(i) $\nabla \times \vec{H} = \vec{J}$	(i) $\oint \overrightarrow{H.dL} = \vec{J}.\vec{dS} = Iconduction$
(ii) $\nabla \times \vec{E} = 0$	(ii) $\oint \vec{E}.\vec{dL} = 0$
(iii) $\nabla . \vec{D} = \rho_v$	(iii) $\oint \vec{D}.\vec{dS} = \oint_{vol} dv$
(iv) $\nabla.\vec{B} = 0$	(iv) $\oint \vec{B}.\vec{dS} = 0$

Maxwell's equations for time variant electromagnetic field:-

(i) $\nabla \times \vec{H} = \vec{J} + \frac{\partial D}{\partial T}$	(i) $\oint \vec{H}.\vec{dL} = \oint_S \left(J + \frac{\partial D}{\partial t} \right).d\vec{S} = I_{total}$
(ii) $\nabla \times \vec{E} = -\frac{\partial B}{\partial T}$	(ii) $\oint \vec{E}.\vec{dL} = \oint \frac{\partial B}{\partial t}.d\vec{S}$
(iii) $\nabla.\vec{D} = \rho_v$	(iii) $\oint \vec{D}.\vec{dS} = \oint_{vol} \rho_v dv$
(iv) $\nabla \vec{E} = 0$	(iv) $\oint \vec{B}.\vec{dS} = 0$

Phaser:- The phaser in the electromagnetic field is analogous to the logarithm in the number field. As the log simplifies the calculation for numbers, the phasor simplifies the calculation for numbers , the phasor simplifies the calculation involved in the analysis of E.M waves. The phasor is therefore a tool which makes a life easier in electromagnetic.

If the expression for time varying x-component of electric field E_x is

$$E_x = E_{x0}Cos(Wt + \psi)$$

Then, It's phasor representation is given by,

$$E_{xs} = E_{xo}e^{j\psi}$$

To recover the original Ex from E_{xs} (like taking antilog in number field). We multiply, E_{xs} by e^{jwt} and then we take the real part i.e.

$$e^{j\psi t} . E_{xo} = E_{xo}e^{j\psi} .e^{j\psi}$$

$$= E_{xo} e^{j(\psi+wt)}$$

$$= E_{xo}[Cos (\psi+wt) + Sin(\psi+wt)]$$

$$R_e[e^{jwt}E_{xs}] = E_{x0}Cos(\psi+wt)$$

Which is the original equation.

The phasor simplifies the differenciation of Ex as shown below:-

We know,

$$E_x = E_{xo}Cos (\psi+wt)$$

Then, $\frac{\delta Ex}{dt} = -WE_{xo}Sin (\psi+wt)$

but, $Re[jwE_{xs}]e^{jwt} = Re [E_{xo}We^{j(\psi+wt})$

$$= Re[E_{xo}[e^{jwt}[Cos(wt +\psi) +jsin(\psi+wt)]$$

$$= Re[E_{xo}[jw[Cos(wt +\psi) - wsin(\psi+wt)]$$

$$= E_{xo} (- wsin(\psi+wt))$$

$$= WE_{xo} sin(\psi+wt)$$

Thus the multiplication of phasor quantity by $jw.e^{jwt}$ and taking it's real par t is equivalent to the differentiation of that quantity in time domain. Therefore the phasor has replaced the complex differentiation with the simple multiplication.

The wave equation:- The wave equations in E.M field (for electric field) is represented as follows:-

$$E_x = E_{xo} Cos[W(t - z\sqrt{\mu E})]$$

Where, $\sqrt{\mu E}$ =1/V = 1/Fλ= T/λ is the phase velocity of the E.M wave in a medium having permeability and permittivity .

Direction of travel:-

We have,

$$E_x = E_{xo} \text{ Cos}[W(t - \sqrt{\mu E})]$$

$$= E_{xo} \text{ Cos}[-(Wz\sqrt{\mu E} - wt)]$$

$$= E_{xo} \text{ Cos } [\frac{2\pi}{T}.z\frac{T}{\lambda} - \frac{2\pi}{T}.t]$$

$$= E_{xo} \text{ Cos } [\frac{2\pi}{\lambda}.z - \frac{2\pi}{T}.t]$$

Now,

For t = 0

For t = 0

$$E_x = E_{xo} \text{ Cos } [\frac{2\pi}{\lambda}.z]$$

For, t = T/4

$$E_x = E_{xo} \text{ Cos } [\frac{2\pi}{\lambda}.z - \frac{\pi}{2}]$$

$$= E_{xo} \text{ Sin } [\frac{2\pi}{\lambda}.z]$$

For, t = T/2

$$E_x = E_{xo} \text{ Cos } [\frac{2\pi}{\lambda}.z - \pi]$$

$= E_{xo}$ Cos $(\frac{2\pi}{\lambda}.z)$

From the figure above it is clear that, a point 'P' which is under consideration on the wave is traveling in z direction .

Velocity:-

We have,

$E_x = E_{xo}$ Cos(W(t-z $\sqrt{\mu E}$))

The quantity (W(t-z $\sqrt{\mu E}$) is constant for any point on the wave provided the distance of the point is measured from a certain reference line. We assure the reference line at the point where for t = 0 , z = 0 . Thus, for t = 0 , z = 0

(W(t-z $\sqrt{\mu E}$) = 0

Again for,

T = T/4, Z = λ/4

(W(t-z $\sqrt{\mu E}$)

=W (T/4 - λ/4. T/λ)

= 0

Again for,

T = T/2, z = λ/2

(W(t-z $\sqrt{\mu E}$) = 0

And so on,

The quantity (W(t-z $\sqrt{\mu E}$) is therefore a constant and is equal to zero.

$\therefore$(W(t-z $\sqrt{\mu E}$) = 0

t-z $\sqrt{\mu E}$ = 0

z = t/$\sqrt{\mu E}$

or, z/t = 1/$\sqrt{\mu E}$

or, dz/dt = 1/$\sqrt{\mu E}$

Here, dz/dt is the rate of change of distance w.r.t time, which is a velocity subtracting the values for air or free space i.e, $\mu_o = 4\pi \times 10^{-7}$

$\varepsilon_o = 4.85 \times 10^{-12}$ F/m

$$\therefore \frac{dz}{dt} = \frac{1}{\sqrt{\mu_o \in_o}} = \frac{1}{\sqrt{4\pi \times 10^{-7} \times 8.854 \times 10-^{12}}}$$

$\approx$ 3 x 10^8 m/s

Which is equal to the velocity of the light. The E.M. wave therefore travels with a velocity of light in free space or air.

Wave impedance:-

The wave impedance is defined as

$$\text{Wave impedance} = \frac{\text{Electric field component}}{\text{Magnetic field component}}$$

For TEM(transverse Electromagnetic wave) There exists only one component of each of the electric and magnetic fields, therefore we have only one wave impedance which is generally known as intrinsic impudence i.e.

$$\eta = \frac{E_x}{H_y} \Omega$$

Where,

η = Intrinsic impedance

E_x = X – component of EF

H^y= Y-component of MF

For Lossy medium i.e. 0 < 6 < ∞

$$\eta = \sqrt{\frac{jw\mu}{\sigma + jw\in}} = \eta_m < \theta_n$$

For perfect dielectric i.e.

$$\eta = \sqrt{\frac{\mu}{\in}}$$

Here, σ = conductance (

W = angular freq. (rad/sec)

Poynting vector:-

A pointing vector $\vec{P}$ is the cross product of $\vec{E}$ (electric field) and $\vec{H}$ (magnetic field). i.e.

$$\vec{P} = \vec{E} \times \vec{H}$$

The magnitude of represents the instant aneous power density (w/m2) at a point and it's direction indicated the direction of the power flow at that point and it's perpendicular to the plane containing $\vec{E}$ and $\vec{H}$

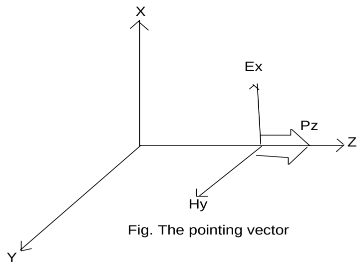


Fig. The pointing vector

Case I:- TEM move traveling in a perfect dielectric (i.e. $\sigma = 0$) we know the wave equation of electric filed is given by;

$$E_x = E_{xo}\mathrm{Cos}[w(t - z\sqrt{\mu \in})]$$

$$= E_{xo}\mathrm{Cos}[wt - \beta z]$$

For perfect dielectric,

$$\eta = \frac{E_x}{H_y} = \sqrt{\frac{\mu}{\in}}$$

$$\text{Or, } H_y = \frac{E_x}{\eta} = \frac{1}{\eta} E_{xo} Cos[wt - \beta z]$$

$$\therefore P_z = E_x H_y = \frac{1}{\eta} E^2{}_{xo} Cos^2[wt - \beta z]$$

Case II:- TEM wave traveling in a lossy medium. The wave in a lossy medium is found to be exponentially attened as distance z increases. But the degree of attemation depends upon the medium itself and is given by which is called attennation constant. The wave equation in this case is given by

$$E_x = [E_{xo}\mathrm{Cos}(wt - \beta z))\, e^{-\alpha z}$$

Where, $e^{-\alpha z}$ = attenation coefficient.

For lossy medium, $\frac{E_x}{H_y} = \eta_n < \theta_n = \eta_m e^{i\theta_n}$

$$= \eta_m e^{j\theta}$$

$$\therefore H_y = \frac{E_x}{\eta_m\, ej\theta} = \frac{1}{\eta_m} E_x e^{-j\theta}$$

$$= \frac{1}{\eta_m} e^{-z} E_{xo} Cos[wt - \beta z] e^{-j\theta}$$

$$H_y = \frac{1}{\eta_m} e^{-(\alpha z + j\theta)} E_{xo} Cos[wt - \beta z]$$

$$H_{ys} = \frac{1}{\eta_m} e^{-(\alpha z + j\theta)} E_{xo}.e^{-j\beta z}$$

$$H_{ys} = \frac{1}{\eta_m} e^{-\alpha z} E_{xo} e - j(\beta z + \theta)$$

The above equation is represented in phasor form in order to recover. Original equation. We go in the following manner.

$$\therefore H_y = \mathrm{Re}\left[\frac{1}{\eta_m} e^{-\alpha z} E_{xo} e^{-j(Bz+\theta)}.e^{(jwt)}\right]$$

$$= \frac{1}{\eta_m} E_{xo} e^{-\alpha z} Cos[wt - \beta z - \theta]$$

The magnitude of the paynting vector is given by

$P_z = E_x. H_y$

$$= \frac{1}{\eta_m} e^{-2\alpha z} E^2{}_{xo} Cos(wt - \beta z).Cos(wt - \beta z - \theta)$$

$$= \frac{1}{2\eta_m} e^{-2\alpha z} E^2{}_{xo}[Cos(2wt - 2\beta z - \theta).Cos\theta]$$

The average P_z is given by,

$$P_{z,\ avg} = \frac{1}{2\eta_m} E^2{}_{xo} e^{-2\alpha z} Cos\theta$$

Retarted potential :- The scalar electric potential at a point caused by a linear charge density is defined as,

$$V = \frac{\rho_L dL}{4\pi \in r}.(V) \_\_\_\_\_\_\_\_\_\_\_\_\_\_ (i)$$

Where r is the distance between dL and the point of interest.

Similarly, the vector magnitude potential is defined as,

$$\vec{A} = \int \frac{\mu Id\ \vec{L}}{4\pi \in r} \quad (wb/m) \ \_\_\_\_\_\_\_\_\_\_\_\_\_\_ (ii)$$

The direction of $\vec{A}$ is same as that of the current. In above equation (i) and (ii), ρ_L and I do not change with time and therefore v and $\vec{A}$ at the point of interest are fixed for all the time. But if and I vary with the time then their values seem at the time of measurement cannot be used to calculate the V and $\vec{A}$ at a distant point. Because it takes time to reach the effect from the source to the point of interest, the values of ρ_L and I which actually contributed the effects have therefore already been changed to same other new value.

Therefore, the above equation are modified as follows:

$$\vec{V} = \int \frac{[\rho_L]d\vec{L}}{4\pi \in r}$$

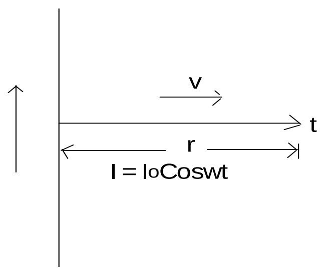
And $\vec{A} = \int \frac{\mu[I]d\vec{L}}{4\pi r}$

The V and $\vec{A}$ in above equation are respectively termed as retarted scalar electric potential and retarted vector magnetic potential . The symbol [ ] represents that the corresponding quantity has been retarted in time in order to compensate the time elapsed in propagating the effect from the source to the points where two quantity is being calculation.

The sketch in fig (i) shows the effect propagating with the velocity of V from the source carrying the current I to the point of intersect P at a distance r. The retarted current in this case is given by,

$[I] = I_o CosW(t – t')$

$[I] = I_o CosW(t –r/w)$



Polarization:-

As the electromagnetic wave propagates the electric and magnetic field components changed behavior is termed as polarization when the direction F is not stated the polarization is taken to be the polarization in the direction of the maximum gain. In practice polarization of the radiated energy varies with the direction from the centre of the antenna so that different paths patterns may have different polarization. There are 3 types of polarization.

(a) Eleptical (ii) Circular (iii) Linear.

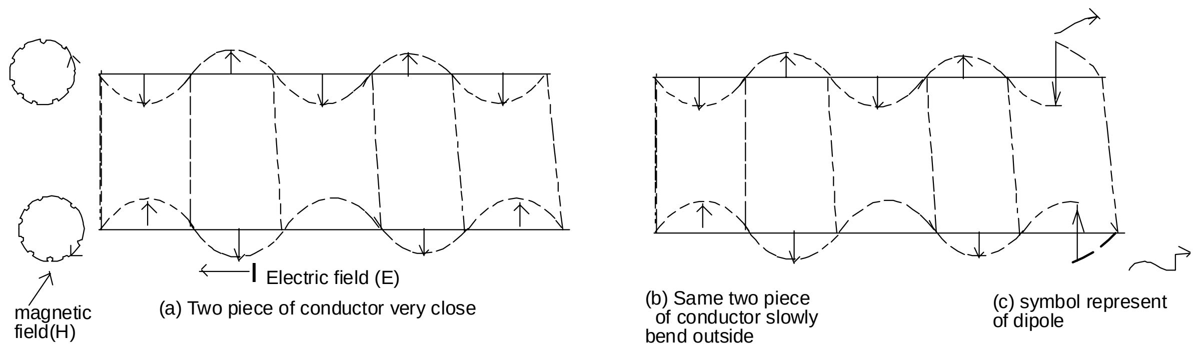
The linear and circular polarizations are the special cases of elliptical polarization.

Linear antenna:- These are broadly categorized into 2 parts:-

(i) Standing wave linear antenna dipole.

(ii) Traveling wave linear antenna dipole.

Standing wave linear antenna:-



It is used for specific frequency because it's i/p impedance is reactive is highly sensitive to frequency. When a piece of open X'mission line is considered there exist a standing wave as illustrated in fig (i)[a] because the conductors are very close to each other. The fields i.e (EF and EM) produced by the individual conductor therefore cancel with each other. Hence, there will not be any radiation from the live. However, if a portion of the line at the open end is slowly bent out – ward the cancellation of the fields decreases gradually when the line finally takes the form as shown in the fig 1[b] no more cancellation occurs and the construction radiated EM waves out into the surrounding medium. The portion of the line which has been bent is the standing wave linear antenna and is popularly known as dipole. There are 3 types of dipole.

(i) Infinite small dipole.

(ii) Short dipole.

(iii) Long dipole.

(i) **Infinite small dipole:-**

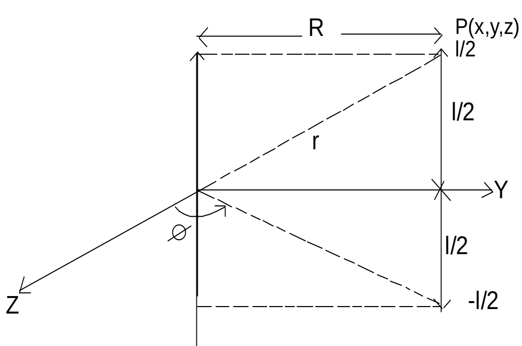


Fig. AN infinite small dipole in carterian co-ordinate system

These type of antennas are building block of practical linear antennas. It has length $l < \lambda/50$ and carries the current $I = I_o Cos(wt - Bz)$, But since the length of the antenna is very small, it is assumed that the current is very uniform through out the length at any time.

At point p(x, y, z)

$$\vec{A}(x,y,z) = \int \frac{\mu[I]d\vec{L}}{4\pi R} = \int \frac{\mu I_o Cos(wt - R/V)d\vec{L}}{4\pi R}$$

$$= \int \frac{\mu I_o Cos(wt - BR)d\vec{L}}{4\pi R}$$

Where, $B = 2\pi/\lambda$, Phase constant

$$\vec{A}_s(x,y,z) = \int \frac{\mu I_o e^{-jBR}}{4\pi R}.d\vec{L} \approx \int \frac{\mu I_o e^{-jBR}}{4\pi R}$$

And also, $\vec{A}_s = \frac{\mu I_o e^{-jBR}}{4\pi R}.l.\vec{a_z}$

We have,

$$A_{rs} = \frac{\mu I_o l e^{-jBR}}{4\pi R}.Cos\theta$$

$$A_{\phi s} = -\frac{\mu I_o l e^{-jBR}}{4\pi R}.Sin\theta$$

$$A_{\phi s} = 0$$

Now, using the relation,

$$\vec{H} = \frac{1}{\mu}\nabla \times \vec{A}$$

$$\vec{Hs} = \frac{1}{\mu}\nabla \times \vec{A_s}$$

The magnetic field component can be calculated as follows:-

Hrs =

Hqs = 0

$$H\phi s = \frac{jBI_o l Sin}{4\pi r}\theta(1 + \frac{1}{JBr})e^{-jBr}$$

Again using the relation

$$E = \frac{1}{JWE}.\nabla \times \vec{H}$$

We get,

$$E_{rs} = I_o L Cos\theta\left(1 + \frac{1}{JBr}\right)e^{-jBr}$$

$$E_{QS} = J\eta \frac{BIoLSin\theta}{4\pi r}\left(1+\frac{1}{JBr}-\frac{1}{(Br)^2}\right)e^{-JBr}$$

Where, η = intrinsic impedance.

It is found that the electric field and the magnetic field components change with a certain fashion within certain range of distance 'r' measurecd from the antenna to the point of interest. Therefore, The space surrounding the antenna is divided into three regions accordingly to make the calculation of the field components easier, They are:-

(a) Near or reactive field region(r<<d)

(b) Intermediate or fresnel field region (r>d)

(c) For or fraunhofer field region (r<<d or 1/Br>>1).

The region in which the distance of the point of interest r, from the antenna is very small in comparing to the operating wave length is called near or reactive field region, where r << d or 1/Br >>1. In this case the above, original equations reduces to the following forms.

$$\therefore E_{rs} = \frac{\eta I_o LCos\theta}{4\pi r^2}\left(\frac{1}{JBr}\right)e^{-jBr}$$

$$= \frac{J\eta I_o LCos\ \theta}{4\pi Br^2}e^{-j\beta r}$$

$$== \frac{J\eta I_o Le^{-jBr}}{4\pi Br^3}Cos\ \theta$$

$$E_{\theta s} = \frac{J\eta\beta I_o LSin\theta}{4\pi r}\left(\frac{1}{\beta^2 r^2}\right)e^{-j\beta r}$$

$$= \frac{-J\eta I_o Le^{-j\beta r}}{4\pi\beta r^3}Sin\theta$$

And $E_{\phi s} = 0$

Also, $H_{\phi s} = H_{\theta s = 0}$

$$H_{\phi s} = \frac{J\beta I_o LSin\theta}{4\pi r}\left(\frac{1}{J\beta\ r}\right)e^{-j\beta r}$$

$$= \frac{I_o Le^{-j\beta r}}{4\pi\beta r^2}Sin\theta$$

Also, we should know that the average power density is given by ,

$$P_{av} = \vec{E}_s \times \overrightarrow{H_s}$$

$$= \frac{1}{2}\mathrm{Re}[\vec{E}_s \times \overrightarrow{H_s}] = 0$$

This shows that there is no power following within the near field region rather electric field changes to magnetic field forms and vice versa instead of propagating.

(b) **Intermediate or fresnel field region (r >d or 1/Br <1):**

The region in which the distance of the point of interest r, from the antenna is greate4r that the operating wavelength is called intermediate field region, where R> d, i.e 1/Br<1. The original equation in this case take the following forms.

$$E_{rs} = \frac{\eta I_o L e^{-jBr}}{4\pi r^2} Cos\theta$$

$$E_{rs} = = \frac{-J\eta I_o L e^{-j\beta r}}{4\pi\beta r} Sin\theta$$

$$E_{\phi s} = 0$$

$$H_{\phi s} = \frac{J\beta I_o e^{-j\beta r} L}{4\pi r} Sin\theta$$

$$H_{rs} = H_{rs} = 0$$

And $P_{av} = \frac{1}{2}\mathrm{Re}[\vec{E}_s \times \vec{H}_s] = \frac{\eta}{2}\left[\frac{\beta I_o L}{4\pi r} Sin\theta\right]^2 \vec{a}_r$

© Far or fraunhofer field region (r <<d or 1/Br <<1):-

The region is which the distance of the point of interest r, from the antenna such that r>> d or 1/Br <<1 is called the far or fraunafor field region. In this case, the equations can take the following forms.

$$E_{\theta s} = \frac{-J\eta I_o L e^{-j\beta r}}{4\pi r} Sin\theta$$

$$E_{rs} \cong 0 \quad E_{\phi s} = 0$$

$$H_{\phi s} = \frac{\beta I_o L e^{-j\beta r}}{4\pi r} Sin\theta$$

$$H_{rs} = H_{\theta s} = 0$$

And Pav = $\frac{1}{2}\mathrm{Re}[\vec{E}_s \times \vec{H}_s] \rightarrow exists$

And generally assumed in antenna.

(iii) **Short dipole:-** A short dipole antenna.

(a) has length, which satisfy 1/50 <l<

(b) Possess current distribution as:-

$$I = \begin{cases} I_o(1-\frac{z}{l}z)Coswt.\overrightarrow{a_z} \\ I_o(1-\frac{z}{l}z')Cos \end{cases}$$

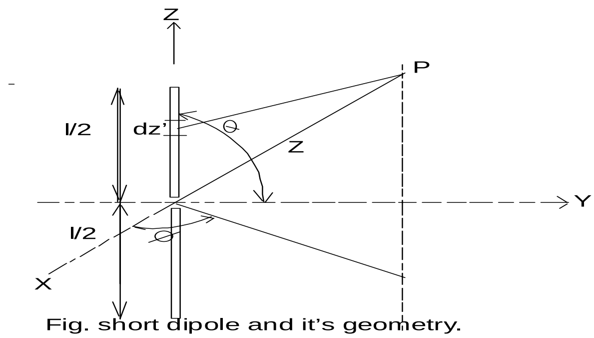


Fig. short dipole and it's geometry.

From fig(i), the vector magnetic potential for the small element dL = dz') can be given by,

$$\vec{A}(x,y,z) = \int \frac{\mu[I]}{4\pi R}dL$$

or, $\vec{A} = \int_{-l/2}^{0} \frac{\mu}{4\pi R} I_o(1+\frac{z}{l}z')Cos[w(t-R/v)]dz' + \int_{0}^{l/z} \frac{\mu}{4\pi R} I_o(1+\frac{z}{l}z')Cos[w(t-R/v)]dz'.\overrightarrow{a_z}$

Also, we can write,

Cos[w(t-R/V)]

= Cos(wt - βR)

$$\therefore \vec{A} = \int_{-l/2}^{0} \frac{\mu}{4\pi R} I_o(1+\frac{z}{l}z')Cos[wt-\beta R)]dz' + \int_{0}^{l/z} \frac{\mu}{4\pi R} I_o(1+\frac{z}{l}z')Cos[w(t-\beta R)]dz'.\overrightarrow{a_z}$$

$$\overrightarrow{A_s} = \frac{\mu}{4\pi R} I_o e^{-J\beta R}\left[\int_{-l/z}^{0}(1+\frac{z}{l}z')dz' + \int_{0}^{l/z}(1-\frac{z}{l}z')dz'\right].\overrightarrow{a_z}$$

Or, $= \frac{\mu I_o e^{-J\beta R}}{4\pi R}\left[z' + \frac{z}{l}\frac{z^2}{2}\right]_{-l/2}^{0} + \left[z' - \frac{z}{l}\frac{z'^2}{2}\right]_{02}^{l/2}.\overrightarrow{a_z}$

$$= \frac{\mu I_o e^{-J\beta R}}{4\pi R}.\overrightarrow{a_z}$$

we consider,

$\therefore$ As $= \frac{-\mu I_o Zl}{16\pi r}e^{-J\beta R}$

Thus we can write, $\vec{E}$ & $\vec{H}$ fields radiated by small dipole (shot dipole) by using.

$$\vec{H}s = \frac{1}{\mu}\nabla \times \vec{A}c$$

$$\vec{E}s = \frac{1}{Jw \in}\nabla \times \vec{H}s$$

$$E_{\theta s} \approx \frac{1}{2}\frac{\left(J\eta\beta I_o zle - j\beta r\right)}{4\pi r}Sin\theta$$

$$E_{rs} = E_{\phi s} = 0$$

&

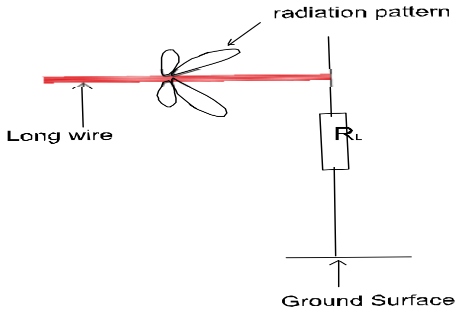
$$H_{\phi s} = \frac{1}{2}\left(\frac{J\eta\beta I_o zle - j\beta}{4\pi r}Sin\theta\right)$$

$$H_{rs} = H_{\theta s} = 0$$

(iii) **Long dipole (see yourself):-**

(2) Traveling wave linear antenna:- The traveling wave antenna can be used over a band of frequencies because it's input impedance is resistive that is not sensitive to the frequency. Since the transmission line or the metallic conductor serving as an antenna will be terminated with it's characteristic impedance, there would be only traveling wave and no more reflecting waves, hence the name traveling wave linear antenna.

(i) **Long wire antenna:-**



A long wire antenna has length, l >>d. It is assumed to be lossless, i.e, the constituent material is lossless and carries the current which is distributed as:

$$\vec{I} = I_o Cos(Wt - \beta z)\ \vec{a_z}$$

The electric and magnetic fields can be estimated as:

$$E_{\theta s} = \frac{J\eta\beta I_o l e^{-J\beta r}}{4\pi r} ej(\beta l/2)(k - Cos\theta) Sin\theta \frac{Sin[\beta l/2)(Cos\theta - k)}{\beta l/2(Cos\theta - k)}$$

$$H\phi s = \frac{E\theta s}{\eta}$$

$$E_{rs} = E_{\phi s} = H_{\theta s} = H_{rs} = 0$$

Here, $k = \frac{\beta w}{\beta}$

$$= \frac{\text{Phase constant along the wire}}{\text{Phase constant in free space}}$$

Characteristics:-

(i) The lobes, which are near the axis of antenna in the direction of the wave is the largest and is called the major labes.

(ii) The pattern is not symmetrical about the axis $\theta = 90^0$

(iii) The terminating resistance is given by

$R_L = 138 \log_{10} (4h/d)$

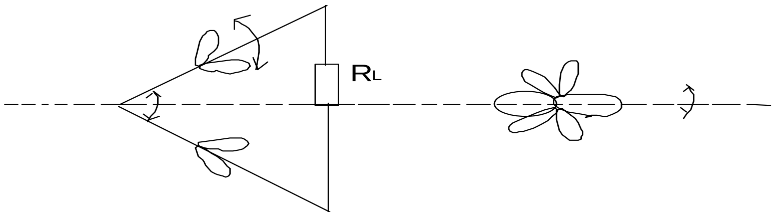
Where, h = height of antenna.

d = diameter of the antenna.

(iv) Simple economical & cost effective.

(v) It can be used to transmit and receive the wave in MF (300Khz – 3MMz) & HF(3 – 30MHz) ranges.

(2) **V – antenna:-**



In case of long wire antenna increasing the length of dipole, the directivity can be increased (i.e it's major lobe towards the direction of propagation of wave). But labes start to split as seen as the length exceeds the operating wavelength. This draw book could be overcome by using a V – antenna. In V –antenna, the number of labes also increases as the length of the wire increases. But the minor labes can be reduced by properly adjusting the angles $2\theta_o$ and $2\theta m$.

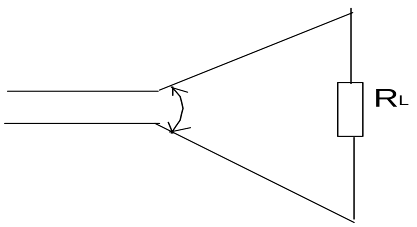


Fig. (i) practical V- antenna

Characteristics:-

(i) $2\theta_o = 2\theta m$, the pattern of each leg of the antenna add in the direction of the line bisecting the antenna, 2θ and forms on major labe in the plane of V.

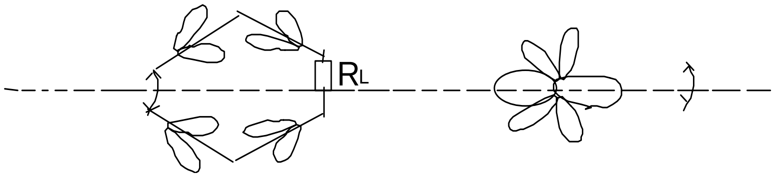
(ii) When $2\theta_o > 2\theta m$, the major labe splits into two distinct labes.

(iii) When $2\theta_o > 2\theta m$, the pattern will be the same as that obtained with $2\theta_o > 2\theta m$ but tilts upward from the plane of the V.

(iv) The value of $2\theta_o$ is calculated as:-

$$2\theta_o = \begin{cases} -149.3(l/\lambda)^2 + 603.4(1/\lambda)^2 - 809.5(1/\lambda) + 443.6 & \text{for } 0.5 \le 1/\lambda \le 1.5 \\ 13.39(1/\lambda)^3 - 78.27(1/\lambda) + 169.77, & 1.5 < 1/\lambda \le 3 \end{cases}$$

(3) **Rhombic Antenna:-**



A rhombic antenna is form by connecting two v-antennas at their ends. The value or RL is equal to the open end characteristic impedance of the V-wire transmission line. The pattern cab be controlled by varying the element length, angle between the element and the plane of the thombous. The advantages of the rhombic antenna over the v-antenna is that it is less difficult to terminate because the ends are nearer to each other then those in the V-antenna.

Antenna Theorem:-

The following are the commonly used antenna theorems in the analysis of any type of antenna:-

(i) **equality of directional patterns:-**

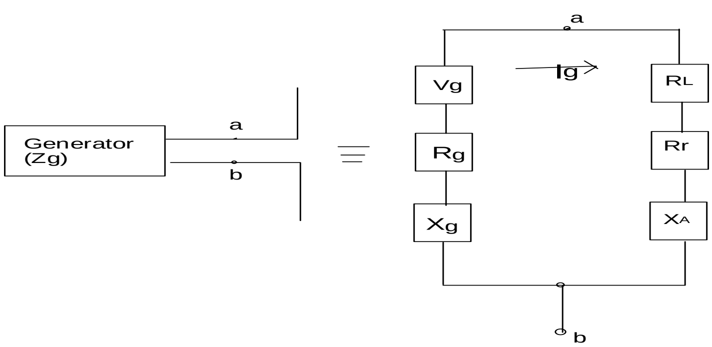
The direction al patterns of receiving antenna is identical with it's directional patterns as transmitting antenna.

(ii) **Equivalence of transmitting and receiving antenna impedance:-**

The impedance of an antenna when used for receiving is the same as when use for the transmitting.

(iii) **Reciprocity:-** If 'I ' is the current apply to the antenna '1' and 'V' is the induced voltage into the antenna '2' then the same voltage 'V' will be observed in the antenna '1' of the current apply to the antenna '2' is equal to 'I'.

(iv) **Thevenin's Theorem:-**



An antenna system can be resolved into it's thevenins' equivalent ciccuit for transmitting antenna.

Here,

R_L = Loss resistance of the antenna.

R_r = Radiatio resistance of the antenna.

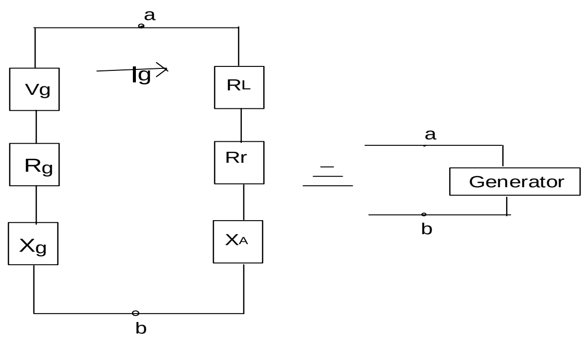
X_A = Reactance of the antenna.
R_g = resistance of the generator
X_g = Reactance of the generator
V_g = Voltage of the generator
I_g = equivalent loop current in the circuit.

(v) **Maximum power transfer.**

An antenna absover or transmit the maximum power from the source when it's impedance is equal to the conjugate of the impedance seen looking back into the source i.e. $R_L + R_r = R_g$

and $X_A = - X_g$

(vi) **Compensation Theorem:-**



An antenna may be replaced by a generator of zero internal impedance whose generated voltage at every instant is equal to the instantaneous potential difference that exists across the antenna.

(vii) **Superposition Theorem:-**

The field intensity at a point due to the number of transmitting antenna is equal to the vector sum of the field intensity at that point due to each of the antennas.

If
$\vec{F}_1$ = field intensity due to antenna r to 1.
$\vec{F}_2$=Field intensity due to antenna r to 1.
$\vec{F}_3$=field intensity due to antenna r to 1.
$\vec{F}_4$=field intensity due to antenna r to 1.
.
.

$\vec{F}_n$=field intensity due to antenna r to 1.

Then, The overall field intensity at point P due to antennas #1, #2, ……….#n is

$$\vec{F}_p = \vec{F}_1 + \vec{F}_2 + \vec{F}_3 + \text{……..} + \vec{F}_n$$

Antenna Fundamental:-

Basic Antenna parameter:-

1. **Radiation Patterns:-**

A graphical representation of the time average power density of the electric and / or magnetic field strengths of an antenna a function o space co-ordinates is known as radiation pattern of the antenna. if the power is plotted the resulting pattern is called the power pattern if the electric or magnetic field strength is taken, the pattern is respectively called the power pattern if the electric or magnetic field strength is taken, the pattern is respectively called the electric or magnetic field pattern.

Forms of radiation patterns:-

(a) Isotropic pattern:-

An hypothetical antenna having equal radiation in all direction is called isotropic pattern, which is independent of θ & ϕ. it is taken as a reference antenna for the study of properties of all typed of practical antenna parameters.

(b) Directional pattern:-

An antenna having the property for radiating or receiving E.M waves more efficiently in some directions then in others is called directional antenna and it’s pattern the directional pattern. It is a function of θ and / or ϕ . Eg. V-antenna, rhombic antenna.

(c) Omni- directional patterns:

An antenna having the property of radiating or receiving E.M waves. As a function of θ (angle of elevation) only is called omni -directional antenna and it’s pattern- omni directional pattern by dipoles.

2. E/H plane:-

The E/H plane is the plane containing the maximum electric/magnetic field vectors. For e.g, X-Z & X – Y planes are respectively the electric and magnetic planes for horn antenna.

3. **Radiation pattern lobes:-**

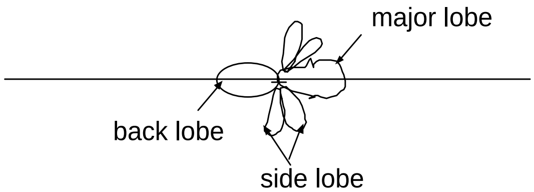


Fig. Basic radiation pattern lobe

A portion of a radiation pattern bounded by the relatively weak radiation intensities is called the radiation pattern lobe. The parts of lobe are:-

i. **Major lobe:-**

It is that portion of radiation pattern which contains the direction of maximum of radiation.

ii. **Minor lobe:-**

Any lobe except the major lobe is called the minor lobe.

a. **Side lobe:-**

The minor lobe located in the hemisphere of main lobe (with or without inclination) is called the side lobe.

b. **Back lobe:-**

The minor lobe located in the hemisphere opposite to that containing the major lobe is called the back lobe.

4. **Radiation Intensity(U):-**

Radiation Intensity in a given direction is defined as the power radiated from an antenna per unit solid angle (steradian) and is obtained simply by multiplying the average power density by the square of the distance i.e

$$U = r^2 pav \quad \text{(W/Steradian)}$$

5. **Directive Gain (D_g) and directivity(θ_o):-**

D_g = radiation intensity of an antenna in a particular direction / Radiation intensity of a reference antenna (an isotropic antenna)

$$D_g = \frac{U_g}{V_o}$$

These quantities show how well an antenna propagates energy in a particular direction . The Radiation intensity U_o is given by

$$V_o = \frac{P_{rad}}{4\pi}$$

$$D_g = \frac{V_g}{P_{rad}/4\pi}$$

$$\therefore D_g = 4\pi \,.\, \frac{V_g}{P_{rad}}$$

And $$D_o = 4\pi .\, \frac{U_{max}}{P_{rad}}$$

6. Total antenna efficiency (e_t):-

The total antenna efficiency is a parameter which indicates the losses at the input terminals and within the structure of an antenna and is calculated as

$$e_t = e_r \, e_{sd}$$

Where,

e_r = reflection efficiency = $1 -1pl^2$

e_{cd} = conduction – dielectric efficiency

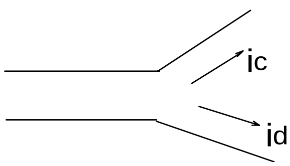
= radiation efficiency

$$= \frac{R_r}{R_r + R_L}$$

R_{er} = The radiation resistance of an antenna

R_L = The loss “ “ “

The loss resistance is used to refer the various conduction and dielectric losses of an antenna. This type of losses are very difficult to compute and hence they are practically measure.



Here,

Ic is the conduction current which occurs in the metal.

$\Rightarrow$ It is the displacement current which is responsible for radiation of EM waves in free space (or in air)

7. Directive power Gain (G_g) and antenna gain (G_0):-

Directive power gain and gain both indicates the directional capability and the efficiency of an antenna and are defined as:-

$$G_g = \frac{4\pi \text{ . Radiation Intensity of an antenna in a given direction .}}{\text{Total input power to the antenna.}}$$

$$= 4\pi .\, \frac{U_g}{\pi}$$

$$= 4\pi \,.\, U_g$$

$$\overline{P_{rad}/e_t}$$

$$= 4\pi\, e_t.\frac{V_g}{P_{rad}}$$

or, $G_g = e_t\,(4\pi\, V_g/P_{rad})$

$\therefore G_g = e_t.\, D_g$

Similarly,

The antenna gain is defined as

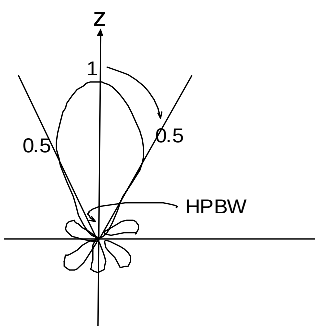
$$G_o = \frac{4\pi.\text{ Maximum Radiation intensity of an antenna}}{\text{Total input power to the antenna}}$$

$$= 4\pi.\frac{U_{max}}{P_i} = 4\pi.\frac{U_{max}}{P_{rad}/e_t} = e_t.\,(4\pi.\frac{U_{max})}{P_{rad}}$$

$\therefore G_o = e_t.\, D_o$

8. Beam wide or half-power beam width(HPOW):-

In a plane containing he direction of the maximum of a beam. The angle between the two directions in which the radiation intensity is one half the maximum value (or $1/\sqrt{2}$) times the maximum value) of the pattern is termed as HPBW.



9. Beam efficiency:-

Beam efficiency is defined as

$$B.E = \frac{\text{Power transmitted within the cone angle } \theta_c}{\text{Total power transmitted by an antenna}}$$

Where,

θ_c =half angle of the cone within which the percentage of the total power is to be found.

or, B.E. = $\int_0^{2\pi}\int_0^{\theta c} U\sin\theta$

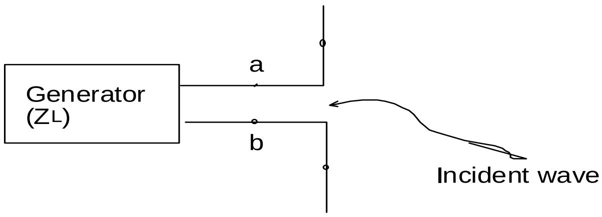
10. **Bandwidth:-**

Band width of an antenna is the range of frequencies within which the performance of an antenna with respect to same characteristic such as input

impedance, the radiation patter, beam width, the antenna efficiency, side lobe ratio and the gain conforms the specified standard.

11. Effective aperture or effective area (Ac) :-

Let us consider the following receiving antenna system:-



It is defined as:

$$A_c = \frac{\text{power delivered to the lead } (P_L)}{\text{Incident power density.}}$$

The effective aperture of an antenna is not necessarily the same as that of physical area or physical aperture. IT is to be noted that a simple wire antenna can capture much more power then it is intersected by it's physical size.

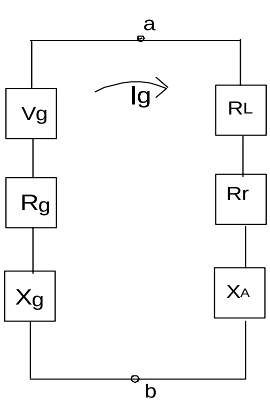
The maximum effective aperture of any type of antenna is related to it's directivity as:-

$$A_{em} = \frac{\lambda^2 D_o}{4\pi}$$

The effective area can also be defined with respect to the transmitting system but we are generally conce4rn about the capture effect characteristic of E.M waves so we are mainly focus on the receiving system.

12. Input Impedance:-

Input impedance is defined as the ratio of voltage to the current measured at the input terminals.



Directional properties of dipole antenna:-

Radio antennas have mainly two functions. The first is to radiate the radio frequency energy that is generated in the transmitter and guided to the antenna by the transmission line. In this capacity the antenna acts as an impedance transforming device to match the impedance of transmission line to that of the free space. The other function of antenna is to direct the energy into the desire direction and it is important to suppress the radiation in other direction where it is not wanted.

For an elementary dipole i.e infinismall dipole having element of Idl, the magnitude of the 'radiation term' for the field strength is given by:-

$$/E/ = E_o = \frac{60\pi IdL}{\lambda r} \times \sin\theta$$

Where, 'θ' is the angle between the axis of dipole and radius vector to the point where strength is measured . we know, for infinitely small dipole and for far field region the field strength is given by.

$E\phi_s = E_{rs} = 0$

$$\& \ E_{\theta s} = \frac{J\eta\beta I_o l e^{-J\beta r}}{4\pi r} Sin\theta$$

But, $\beta = 2\pi/\lambda$

$\eta = 120\pi$ [for very small element $I_o dl = Idl$]]

$$= \sqrt{re/\in_o}$$

$$E_{\theta s} = \frac{j.120\pi \frac{2\pi}{\lambda}.I_o l e^{-j\beta r}}{4\pi r}.Sin\theta$$

$$= \frac{j.60\pi.I_o l e^{-j\beta r}}{\lambda r}.Sin\theta$$

$$\text{or, } /E_{\theta s}/ = \frac{60\pi.I_o l}{\lambda r}.Sin\theta$$

$$\text{or, } /E_\theta/ = \frac{60\pi.Idl}{\lambda r}.Sin\theta$$ [for very small element $I_o dl = Idl$]

which is same as equation (i).

Note:-

- Remaining parts (i.e finding E & H- field studied in 1st chapter).
- Radian pattern of traveling wave antenna also studied in 1st chapter).

Antenna Arrays:-

An antenna array is an assembly or arrangement of radiating element in an electrical and geometrical configuration. The electrical configuration is related with P' the phase among the member element, the geometrical configuration corresponds to the physical parameters such as distance between the element and length the of elements.

The necessity of the antenna arrays arises from the demand of higher directivity required for long distance communication. A single element antenna posses a wide radiation pattern that means low directivity or gain. Enlarging the dimension of single element antenna, it's directivity can also be increased. However this technique might not always work in practice due to various limitation in antenna system be due to the limitation in size, shape, required radiation pattern etc. so, the antenna array comes into play which does do not need to increase the dimension of single element antennas but produce a higher directivity or gain.

In an array of identical element, five controls can be used to save the overall pattern of the antenna.

These are:-

(i) The geometrical configuration of the array system , i.e linear , circular rectangular

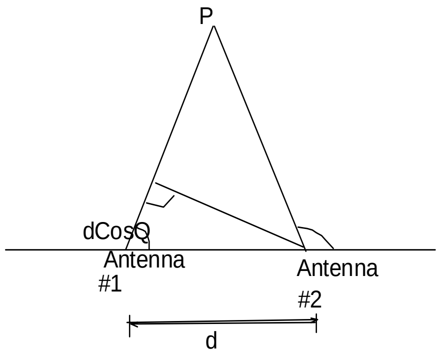
(ii) The relative displacement between the elements.

(iii) The excitation amplitude of the individual element.

(iv) The excitation phase of the individual element.

(v) The individual or (relative) patterns of the elements.

Two element array:-



When greater directivity is required instead of single antenna, antenna array are used. An antenna array is a system of similar antenna similarly oriented. Antenna arrays make use of wave interference phenomena that occur between the radiations from the different element of array.

Consider the two element array in which the antenna # 1 and antenna #2 are situated in the plane under consideration .

Let, r_1 = the distance from antenna #1 to point P.

r_2 = the distance from antenna #2 to point p.

d = distance between the #1 and # 2.

Let, us consider the point P which occur interest to find out the straight of electric field due to antenna #1 and #2. We also consider the point P to be sufficiently remote from the antenna system so that the radius to the point P can be parallel.

From figure (i),

$R_2 = r_1 - d\cos\theta$

But, 'P' point becoming very far from the antenna #1 & @#2.

$r_2 \approx r_1$

$\Rightarrow d\cos\theta \approx 0$

Here,

The path difference in meters is given by

$= d\cos\theta$ (m)

$\therefore$ The path difference in terms of wave length will be $= d/\lambda \cos\theta$

$\therefore$ The phase angle.

$\Psi = 2\pi$ (path difference)

Or, $\Psi = \dfrac{2\pi}{\lambda} \cos\theta$

$\Psi = \dfrac{\beta d}{\lambda} \cos\theta$

Where, β = Phase constant

Also if,

I_1 = the current in antenna # 1

I_2 = "   "   "   #2

Then, we consider,

$I_2 = I_1 < \alpha$

Where, α = the phase angle by which the current I_2 leads I_1

The total phase difference will be,

$\Psi = \beta d \, Cos\theta + \alpha$

Let, E_1 be the electric field strength at point 'P' due to element #1.

E_2 be the electric field strength at point 'P' due to element #2.

Then, total field at point P due to elements #1 and #2 will be.

$$\overrightarrow{E_T} = \overrightarrow{E_1} + \overrightarrow{E_2}$$

We know,

$$\overrightarrow{E_1} = \frac{60\pi Idl}{\lambda r_1} Sin\theta < o \qquad \_\_\_\_\_\_\_\_\_\_\_\_(i)$$

$$\overrightarrow{E_2} = \frac{60\pi Idl}{\lambda r_{21}} Sin\theta < \psi \qquad \_\_\_\_\_\_\_\_\_\_\_\_(ii)$$

Now dividing eq^n (i) by (ii) we get,

$$\frac{\overrightarrow{E_1}}{\overrightarrow{E_2}} = \frac{I_1 < o_1}{I_1 < \psi} \qquad [\because r_1 \approx r_2]$$

$$\frac{\overrightarrow{E_1}}{\overrightarrow{E_2}} = \frac{I_1}{I_2 e^{j\psi}}$$

$$\vec{E}_1 I_2 e^{j\psi} = \vec{E}_2 I_1$$

$$\vec{E}_2 = \vec{E}_1 \frac{I_2}{I_1} e^{j\psi} = \vec{E}_1 k e^{j\psi}$$

Where, $K = \frac{I_2}{I_1}$

$$\therefore \quad \overrightarrow{E_T} = \overrightarrow{E_1} + \overrightarrow{E_2}$$

$$= \overrightarrow{E_1} + \overrightarrow{E_2} K e^{j\psi}$$

$$\overrightarrow{E_T} = \overrightarrow{E_1} + \overrightarrow{E_2}[1 + Ke^{j\psi}] = \overrightarrow{E_1}[1 + K[Cos\psi + J\sin\psi]]$$

$$\overrightarrow{E_T} = \overrightarrow{E_1}[\sqrt{(1 + KCos\,\psi)^2 + k^2 Sin^2}\psi$$

$$= \overrightarrow{E_1}\sqrt{(1 + 2KCos\psi + K^2 Cos^2\psi + k^2 Sin^2}\psi$$

For simplicity, we consider,

$$\therefore \quad \overrightarrow{E_T} = \overrightarrow{E_1}\sqrt{2\,(1 + Cos\quad\psi\,)}$$

$$= \overrightarrow{E_1}\sqrt{2.2Cos^2\psi / 2}$$

$$E_T = \overrightarrow{E_1} 2Cos^2\psi / 2$$

$$\therefore E_T = 2\overrightarrow{E_1} Cos\left[\frac{\beta d Cos\theta + \alpha}{2}\right]$$

$$E_T = 2\overrightarrow{E_1} Cos\left[\frac{\pi}{\lambda} d Cos\theta + \frac{\alpha}{2}\right]$$

Which is the total field strength at point 'P' due to element #1 and #2.

Special Case:-

When $d = \lambda/2$

& $\alpha = 0$

Equation (iii) reduces to

$E_T = 2E_1 Cos[\pi/2 Cos\theta]$

Again we consider,

$2E_1 = 1$

$\therefore\ E_T = Cos[\lambda/2 Cos\theta]$

Now, for maxima,

$Cos[\lambda/2 Cos\theta] = \pm 1$

$\lambda/2 Cos\theta_{max} = \pm n\pi$ for n = 0. 1, 2, ……………

If n = 0 then,

$\lambda/2 Cos\theta_{max} = 0$

$\theta_{max} = 90^0, 270^0$

Now for minima,

$Cos[\lambda/2 Cos\theta_{min}] = \pm (2n + 1)\ \pi/2$ for n = 0, 1, 2, …………..

For n = 0,

$\pi/2\ Cos\theta_{min} = \pm\ \pi/2$

or, $Cos\theta_{min} = \pm 1$

or, $\theta_{min} = 0^0$ & 180^0

for half power point direction,

$Cos[\pi/2 Cos\theta] = \pm 1/\sqrt{2}$

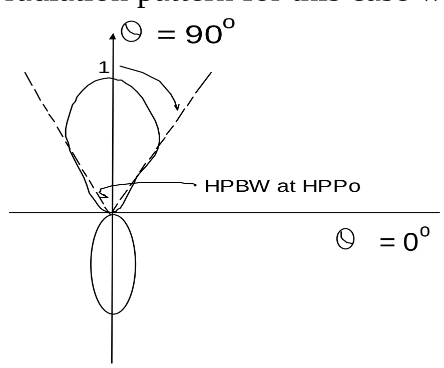
or, $\pi/2 Cos\theta_{HPPD} = \pm (2n + 1)\ \pi/4$

or, $\pi/2 Cos\theta_{HPPD} = \pm\pi/4$ (for n = 0)

$Cos\theta_{HPPD} = \pm 1/2$

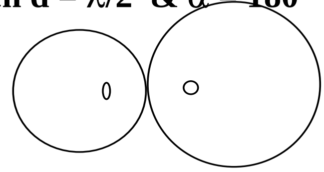
$\theta_{HPPD} = 60^0$ & 120^0

$\therefore$ The radiation pattern for this case will be:-

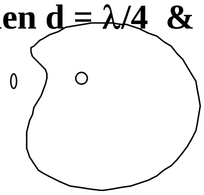


The other forms of radiation pattern for various cases are shown below:-

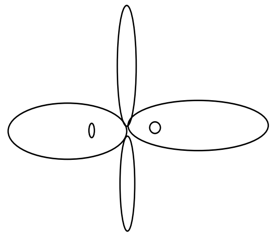
***When $d = \lambda/2$ & $\alpha = 180^0$**



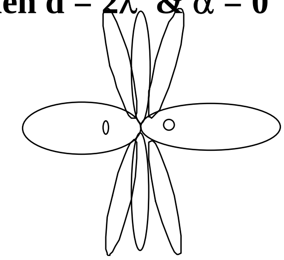
* **When d = $\lambda/4$ & α = -90^0**



* **When d = λ & α = 0^0**



* **When d = 2λ & α = 0^0**



Horizontal pattern in Broadcast array:-

An array having more then two element of array is called broadcast arrays. Since, the pattern produced by the two element array must always be symmetrical about the plane through the antenna and position of an any two nulls can be specified. But in three or more elements array, the antenna configurations and spacing as well as current magnitudes and phases are all variable under the control of the designer, this permit a large number of different antenna pattern types.

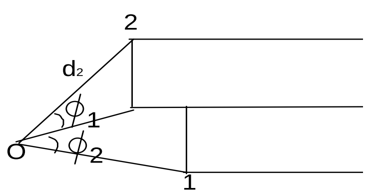


Fig. (i) the simple 3 element arrays system.

For a three element arrays, the resulting pattern is given by,

$E_T = /\in_o \left(1 + K_1 e^{J\psi 1} + K_2 e^{J\psi 2}\right)/$

Where, $\Psi_1 = (\beta d_1) \cos\phi_1 + \alpha_1$

Also, we consider,

$I_1 = I_o < \alpha_1$

$I_2 = I_o < \alpha_2$

For a point to point communication at the higher frequencies the desire radiation pattern is a single narrow lobe or beam. To obtain such characteristic a multi element linear when the elements of the array are spaced equally along a horizontal line. I a linear array the elements are fed with currents of equal magnitude and having a uniform progressive phase shift along the line as shown in fig below.

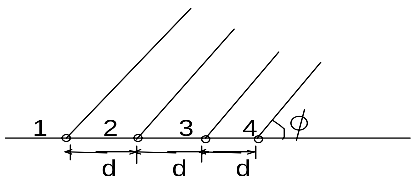


Fig. Linear 4-element array with equal spacing

The total electric field strength can be obtained by adding vertically the field strengths due to each of the elements. i.e.

$$E_T = /\in_o \left(1 + e^{J\psi} + e^{J2\psi} + e^{J3\psi} \ldots\ldots\ldots\ldots\ldots e^{J(n-1)\psi}\right)/$$

Where,

, $\Psi = \beta d \, \text{Cos}\phi + \alpha$ $\quad [\therefore \text{Km} = 1]$

Here,

α is the progressive phase shift between element.

Multiplication of Patterns:-

Statement:-

It can stated as the total field pattern of an array of name isotropic but similar source is the multiplication of the individual source patterns and the pattern of an array of isotropic point source each located at the phase center of individual source and having the resistive amplitude and phase .where as the total phase pattern is the addition of the phase pattern of the individual sources and that of an array of isotropic point source

Here, the pattern of individual source is assumed to be same whether it is in the array or isolated.

Let. E = the total field.

$E_i(\theta,\Psi)$ = field pattern of individual sources.

$E_\alpha(\theta,\Psi)$= field pattern of array of isotropic point source.

$E_{pi}(\theta,\Psi)$ = phase pattern of individual source.

$E_{pa}(\theta,\Psi)$ = Phase pattern of array of isotropic point source.

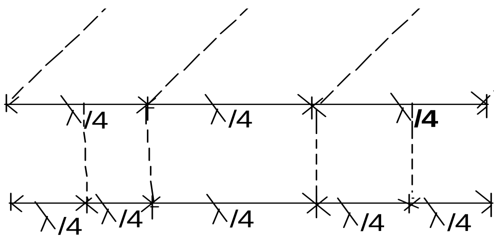
Then,

$$E = \{ E_i(\theta,\Psi) \times E_\alpha(\theta,\Psi) \} \times \{ E_{pi}(\theta,\Psi) + E_{pa}(\theta,\Psi) \}$$

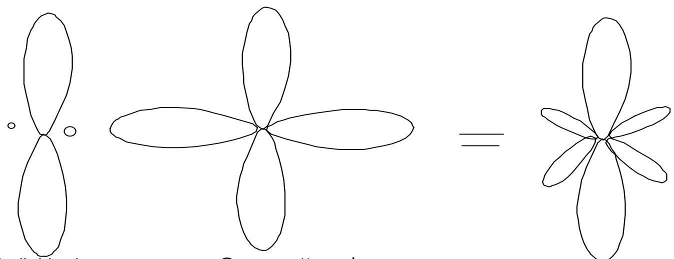
Here, θ and Ψ are called polar and azimuthal angle respectively. The principle of multiplication of pattern provides a speedy method for sketching the pattern of complicated arrays just by the in section and thus the principle proves to be useful tool in the design of an antenna array.

Case:- 1

Radiation pattern of 4- isotropic element fed in phase and spaced by $\lambda/2$



Consider a four element array of antenna in which the spacing between each unit is $\lambda/2$ and currents are in phase (i.e $\alpha = 0$)



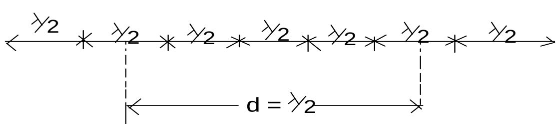
individual source pattern

Group pattern due to array

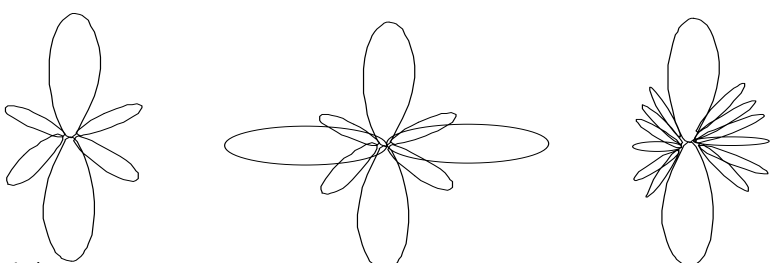
resultant radiation pattern of 4-isotropic element

Case:- 2

Radiation pattern of 8 isotropic element fed in phase and spaced by $\lambda/2$ apart.



The total pattern in this case will be



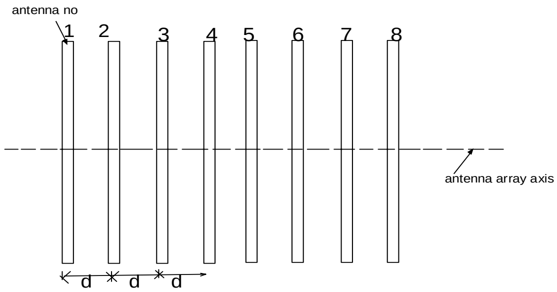
4 element pattern as individual

Group pattern due to array

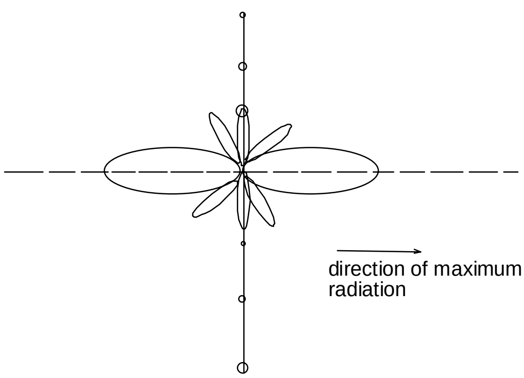
resultant radiation pattern of 8- isotropic element.

Forms of linear antenna arrays:-

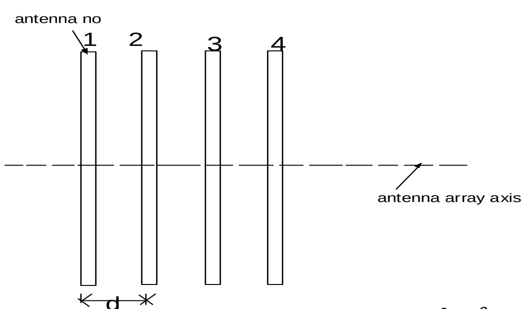
(i) **Broadside array**:-

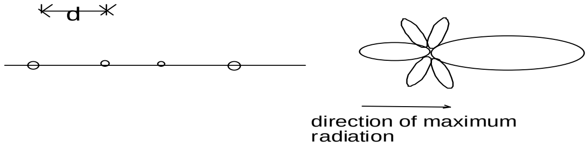


This is one of the important antenna arrays used in practice. Broadside array is one in which a no of identical parallel antennas are set up along a line drawn perpendicular to their respective axis. In the broadside array individual antenna or elements are equally spaced along a line and each element is fed with the current of equal magnitude all in the same phase. By doing so, this arrangement fires in broadside direction (i.e. perpendicular to the line of array axis) where there are maximum radiation s in a particular direction and relatively a little radiation in other directions and hence the radiation pattern of broadside array in brocade sectional. Therefore broadside array may be defined as a"An arrangement in which the principal. Direction is perpendicular to the array axis and also the plane containing the array elements". The general radiation pattern of this type of array is choose below:-



(ii) **End fire array:-**





The end fire array is nothing but broadside array except that individual elements are fed in out of phase (usually 1800). Thus in the end fire array a no of identical antennas are spaced equally along a line and individual elements are fed with currents of equal magnitude but their phases vary progressively along the line in such a way to make the entire arrangement substantially unidirectional. Therefore end fire array may be defined as “An arrangement in which the principal direction of radiation coincides with the direction of an array axis.

Yagi-Uda array:-

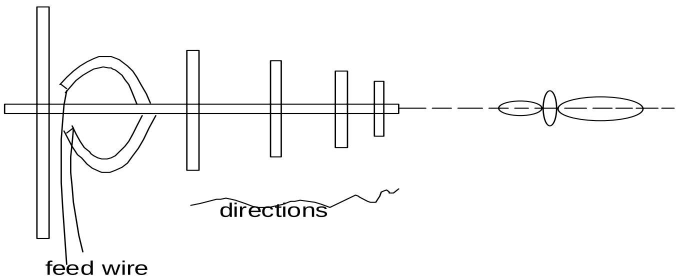


fig. Yagi-uda antenna.

This array consists of folded dipole and a no of parasites elements.. This array is commonly used for HF, VHF & VHF & UHF (3 – 30 mHz, 30-300mHz, 300-3GHz) communications. In this system only the folded dipole is energized directly by a fed transmission line. The parasi. Elements acquire their currents from the mutual induction, that is why these elements are called parasite. The parasitic elements in the direction of the beam are called the direction s while those in the backward direction are called reflectors. To achieve the end-fire beam formation, the directors are made some what smaller in length (about 95% of the drien element) and the reflectors are made some what larger than the folded dipole (about 105% of the folded dipole). The input impedance is purely resistive with the value nearly equal to 300ohm. Which is an ideal match to the commercially available twin lead which also has an impedance of 300 ohm.

The input impedance of folded dipole is calculated from the following relation.

$$Z_{Fd} = 73N^2$$

Where, N is the no of parallel sides having same diameter of dipole for folded dipole, i.e. N = 2.

$\therefore Z_{Fd} = 73 \text{ x } Z^2 = 292 \text{ x } 300$ ohm.

General characteristics:-

(i) If 3 elements array (i.e a reflector a foled dipole and a director) is used than such that type of Yagi-Vda antenna is generally referred to as beam antenna.
(ii) It has unidirectional beam as shown in fig above.
(iii) It provides gain of the order of about 8 dB.
(iv) It is also known as supergain antenna due to it's high gain and high beam width.
(v) IF greater directivity is required further parasitic elements may be used.
(vi) It is essentially a fixed frequency device.

Log periodic array:

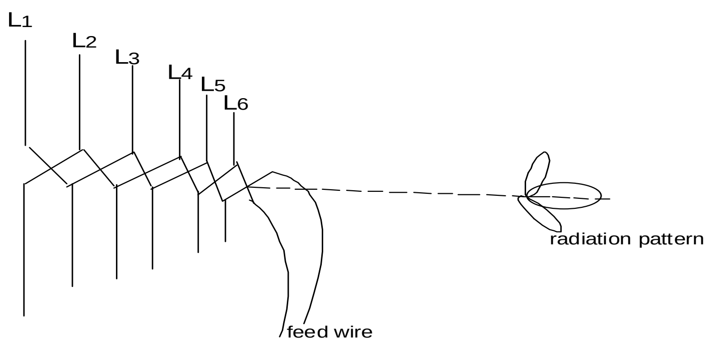


Fig. LOg periodic array

The shortest element is directly fed with transmission line and the succeeding elements acquire both the conduction and induced currents. The former element is the dominant one. The condition current arrives at the succeeding elements later in the time than the proceeding one which causes the phase delay. The separate phase delay also arise from the induced energy. The total effect results in an end fire radiation pattern as shown in fig. (ii).

The log periodic array is capable of operating over a 4:1 frequency ratio such as in OHF(54 mHz- 326 mHz). The input impedance ranges from 200 to 820 ohm. Therefore a impedance matching transformer is usually required to match a log periodic array with the commercially available transmission line.

General characteristics:-

(i) log periodic antenna is excited from the shortest length side.
(ii) For unidirectional log periodic antenna the structure fires in backward direction i.e. towards the shortest element and forward radiation is very small or zero.

(iii) Transmission line in active region must have proper characteristic impedance with negligible radiation.

(iv) In active region current magnitude and phasing should be proper so that strong radiation occur along the backward direction.

Chapter:- 3

Antenna propagation:-

Transmission loss between antenna:-

The transmission loss plays a major tole in the application of radio transmission line. The transmission loss is simply a measure of the ratio of the power transmitted or radiated by an antenna to the power received by the antenna. It is usually expressed in dB.

i.e $$L = \frac{W_T}{W_R} \quad \text{——————(i)}$$

Where, L = the transmission loss.

W_T = transmitted power.

W_R = received power.

Let, us consider the transmission line is in free space. The power density at a distance d from the transmitting antenna is given by,

$$Pd = \frac{W_T}{4\pi d^2} \quad \text{————(ii)}$$

Let, g_1 = Gain due to T_x antenna.

Then, the power density is increased to

$$P_{dinc} = g_1 . Pd = g_1 . \frac{W_T}{4\pi d^2} \quad \text{————(iii)}$$

Let, g_2 = Gain due to receiving antenna, the effective area (or aperature) of the receiving antenna is given by,

$$A = \frac{\lambda_2}{4\pi} g_2, \quad \text{————(iv)}$$

From the definition of effective area,

We know,

$$A = \frac{\text{Received power (i.e power delivered to the load)}}{\text{Instant power density (pd inc)}}$$

$$\text{or, } A = \frac{W_R}{P_{dinc}}$$

$$\text{or, } W_R = A . P_{dinc}$$

$$= \frac{\lambda^2 g_2}{4\pi} . \frac{W_T g_1}{4\pi d^2}$$

$$= W_T g_1 g_2 . \frac{\lambda^2}{\ }$$

$$(4\pi d)^2$$

Hence, the ratio of transmitted power to the received power is given by,

$$\frac{W_T}{W_R} = \frac{(4\pi d)^2}{g_1 g_2 . \lambda^2} \quad \text{\_\_\_\_\_\_\_\_\_\_\_\_\_\_\_(v)}$$

For isotropic antenna,

$$g_1 = g_2 = 1$$

In this case, equation (v) becomes.

$$\frac{W_T}{W_R} = \frac{(4\pi d)^2}{\lambda^2} \quad \text{\_\_\_\_\_\_\_\_\_\_\_\_\_\_\_(v)}$$

$$10\mathrm{Log}_{10} \frac{W_T}{W_R} = 20\mathrm{Log} \frac{(4\pi d)}{\lambda} \quad \text{\_\_\_\_\_\_\_\_\_\_\_\_\_\_\_(vi)}$$

$$L_b = 20\log \frac{(4\pi d)}{\lambda} \text{ dB} \quad \text{\_\_\_\_\_\_\_\_\_\_\_\_\_(vi)}$$

Where, L_b = The basic transmission loss expressed in dB. But the actual transmission loss i.e for nanisotropic cases equation (v) can be written as,

$$10\mathrm{Log}_{10} \frac{W_T}{W_R} = 10\log \frac{(4\pi d)^2}{\lambda^2 g_1 g_2}$$

$$\text{or, } L(\mathrm{dB}) = 20 \mathrm{Log} \frac{(4\pi d)^2}{\lambda} - 10\mathrm{Log}g_1 - 10 \mathrm{Log}\, g_2$$

$$\text{or, } L(\mathrm{dB}) = (L_b - G_1 - G_2) \text{ dB} \quad \text{\_\_\_\_\_\_\_\_\_\_(vii)}$$

G_1 = Gain of T_X antenna in dB = $10\mathrm{Log}g_1$

G_2 = gain of R_X antenna in dB = $10\mathrm{Log}g_2$

Transmission loss as a function of frequency:-

The transmission loss depends upon the circumstances of the problem. Dependability of transmission loss on the basis of operating frequency can be divided for three particular cases:-

Cases:-

(i) **For fixed gain:-** for vehicular communicating air to ground line and navigation (system use both antenna's that have omnidirectional coverage. A monopole or vertical dipole is a typical omni directional antenna. Under this circumference the directional gain is fixed and independent of frequency. Thus from equation (v), we can say that for fixed gain antennas. The received power (transmission loss as well) is proportional to the square of the wavelength which means directly proportional to the operating frequency squared. i.e.

$$\frac{W_T}{W_R} = \frac{(4\pi d)^2}{\lambda^2 g_1 g_2}$$

for fixed gain,

g_1 and g_2 = constant & d = constants.

$$\therefore \frac{W_T}{W_R} \propto \frac{1}{\lambda^2}$$

Which means,

$$\frac{W_T}{W_R} \propto f^2$$

$$L \propto f^2$$

$$\boxed{L_{\alpha B} \propto 20\log f.}\ \text{dB} \quad \text{\_\_\_\_\_\_\_\_\_\_\_(viii)}$$

(ii) **for fixed gain and fixed area antenna:-**

We know from equation (iv), the effective area of receiving antenna is given by

$$A_2 = \frac{\lambda^2.\, g_2}{4\pi}$$

which is fixed for the given case.

Thus,

$$\frac{W_T}{W_R} = \frac{(4\pi d)^2}{\lambda^2 g_1 g_2} = \frac{4\pi d)^2}{g_1} \cdot \frac{4\pi}{\lambda^2 g_2}$$

$$\frac{W_T}{W_R} = \frac{4\pi d^2}{A_2 g_1}$$

$$\frac{W_T}{W_R} \propto \frac{1}{g_1} \quad \text{\_\_\_\_\_\_\_\_\_\_\_(ix)}$$

From the above equation we see that, the loss is proportional to the transmission antenna gain. But in most of the cases to have the uniform omnidirectional pattern g1 is a loss set fixed. So we can say that the loss WT/WR is independent of frequency i.e the loss is fixed. An example may be taken as link between satellite and a ground based antenna, where for simplicity we take satellite as a isotropic radiator.

(iii) **For Microwave links:-**

In case of ordinary microwave links, both the antenna's aer made directional with size, which is limited by cost consideration for this reason, the effective areas for both the transmitting and receiving antennas are given by,

$$A_1 = \frac{\lambda^2 . g_1}{4\pi}$$

And $$A_2 = \frac{\lambda^2 . g_2}{4\pi}$$

Thus, equation (v), reduces to ,

$$\frac{W_T}{W_R} = \frac{(4\pi d)^2}{\lambda^2 g_1 g_2}$$

$$= \frac{(4\pi d)^2}{\frac{A_1 4\pi}{\lambda^2} . \frac{A_1 4\pi . d^2}{d^2}}$$

$$= \frac{(4\pi d)^2 . \lambda^2}{(4\pi)^2 A_1 A_2}$$

$$= \frac{(\lambda d)^2}{A_1 A_2} = \frac{\lambda^2 d^2}{A_1 A_2}$$

$$\therefore L \propto \lambda^2$$

$$LdB \propto 20\log\lambda$$

Which implies that the transmission loss is directly proportional to the operating wave length squared. At the same that, we also say that it is inversely proportional to the square of the operating frequency.

Antenna temperature and signal to nose ratio:-

The signal to noise ratio is an important parameter in the design of a communication channel . The signal to nose ratio is defined as the ratio of the power of the signal to the power of the nose i.e

$$S/N \text{ (or SNR)} = \frac{\text{power of signal}}{\text{Power of noise}} \quad \text{……………………..} \quad \text{(i)}$$

The basic problem in the design of a communication channel is to receive signal strong enough to yield an acceptable SNR. If the noiseness of

the channel (or system) is low the received signal can be very small and still produces a quite adequate channel.

Design of an antenna receiving system requires therefore a knowledge of effects of both antenna nose and receiver nose for satisfactory signal to nose /p. as we known,we usually express the power of nose in terms of watte/Hz of handwith but there is another more convenient measure of nose power in terms of temperature and is given by Nyguist relation. According to this relation, the noise power available from a resistor R at absolute temperature T Kelvin is

$$Pn = KTB \text{ ---------- (ii)}$$

Where, K = Boltzmann's constant = $1.38 * 10^{-23}$J/K

B = Bandwidth in Hz.

The nose power can be expressed in terms of termal nose voltage across R as,

$$P_n = \frac{V^2}{4R} \quad \_\_\_\_\_\_\_\_\_\_\text{(iii)}$$

Where, V = thermal nose voltage.

From equations (i) and (ii)

$$\frac{V^2}{4R} = KTB$$

$$V = 2\sqrt{KTBR} \; \_\_\_\_\_\_\_\_\_\_\text{(iv)}$$

When an antenna is connected to a receiver of gain 'G'. Then both signal and nose are amplified, and in addition noise is added by the receiver so that the input signal to noise ratio is also decreased

If S_A = I/P signal generated by the receiver antenna at a temperature T_A "k then the o/p signal power will be.

$$S = SAG \; \_\_\_\_\_\_\_\_\_\_\text{(v)}$$

The o/p noise power will be the sum of antenna noise and the receiver noise iN. Thus, the o/p noise power 'N' is given by:-

$$N = P_nG + P_NG$$

$$= AT_AGB + kT_eBG$$

$$=K(T_A+T_e)BG \; \_\_\_\_\_\text{(vi)}$$

Where,

T_e = effective noise temperature of the receiver n/w.

Thus, from equation (v) and (vi) the o/p signal to noise ratio is given,

$$SNR = S/N = \frac{S_AG}{K(T_A+T_B)BG}$$

$$\therefore\ SNR = S/N = \frac{S_A G}{K(T_A+T_B)BG}$$

$$\therefore\ SNR = \frac{S_A G}{K(T_A+T_B)B} \qquad \text{\_\_\_\_\_\_\_\_\_\_(vii)}$$

Thus, from eq[n] (vii) we can say o/p signal to noise ratio is proportional to the antenna noise temperature TA the effective noise temperature Te. Thus from the point of view of SNR< for better system design we need to maximize $\frac{S_A}{T_A+T_e}$

Ex:-01

In a satellite communication system, free space condition may be assumed. The satellite is at a height of 3600Km above earth, the freq used is 4 GHz, the X'mitting antenna is 15 dB and the receiving antenna gain is 45 dB. Calculate.

(i) free space transmission loss.

(ii) The receiving power when the transmitted power is 200 watt.

Sol[n]:- Given,

$h = 3600km$

$f = 44Hz$

$g_1 = 15dB$

$G_2 = 45dB$

(i) $L(dB) = ?$

We know,

$L(dB) = L_b - G_1 - G_2 \quad dB$

Where, L = transmission loss dB.

L_b = basic transmission loss in dB.

$$= 20Log\left(\frac{4\pi d}{\lambda}\right)$$

Where,

$$\lambda = c/f = \frac{3 \times 10^8 m/s}{4G_{Hz}} = 0.075m.$$

$$\therefore\ L_b\ (dB) = 20Log\left(\frac{4\pi \times 36000 \times 10^3}{0.075}\right)$$

$$= 195.6dB$$

$\therefore\ L_b(dB) = 195.6\ dB$

$L(dB) = L_b - G_1 - G_2 \quad dB$

(ii) we know,

$$\frac{W_T}{W_R} = L = \frac{(4\pi d)^2}{\lambda^2 g_1 g_2}$$

Where, W_T = X'mitted power
W_R = received power

$g_1 = 10\ G_1/10$

$g_2 = 10G_2/10$

and $L = 10^{L(dB)/10} = 10^{135.6/10}$ =…………….

$\therefore W_R = \frac{W_T}{L} = \frac{200}{10^{135.6/10}}$ ……………..

Free Space propagation:-

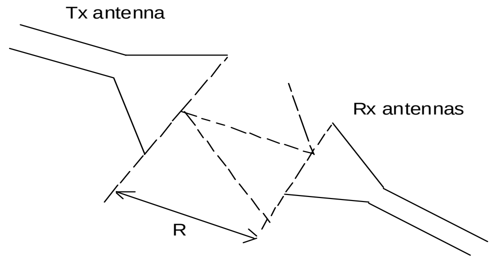


Fig. free space propagation

When a communication between a satellite and ground station is consider. Then the concept of free space propagation comes into existence. In the analysis of free space propagation, the power supplied to the transmitting antenna and power delived to the load of the receiving antenna and their relation are evaluated . For simplicity, the influence of earth and other obstuctals are neglected and it is consider that both the antenna have identical polarization.

Suppose a transmitting antenna is isotropic. Then the time average power density at distance R is defined by:-

$$P_{av}, o, t_x = \frac{P_{rad}. O,.t_r}{4\pi R^2} = \frac{(P_{in}, o, t_x)(e_t, t_x)}{4\pi R^2}$$

Here,The subscript av indicate average, o indicate isotropic, t_x indicate transmitter and t_r indicate receiver.

If we consider the case to be non isotropic, then the directivity from the transmitting antenna is given by:-

$$D_g(\theta_{tx},\phi_{tx}) = \frac{U(\theta_{tx},\phi_{tx})}{U_o\ P_{av},o,t_x}$$

$$= \frac{R^2\ P_{av}(\theta_{tx},\phi_{tx})}{R^2\ P_{av},o,t_x}$$

$$\therefore P_{av}(\theta_{tx},\phi_{tx}) = D_g(\theta_{tx},\phi_{tx})\ P_{av},o,t_x$$

$$= D_g(\theta_{tx},\phi_{tx})\ \frac{P_{in},o,t_x\ (e_t, t_x)}{4\pi R^2}$$

But,

$$P_{in,o,tx} = P_{in,tx}$$

$$\therefore P_{av}(\theta_{tx},\phi_{tx}) = \frac{D_g(\theta_{tx},\phi_{tx})(P_{in,}\ t_x)(e_t,\ t_x)}{4\pi R^2}$$

We know, the effective aperture of receiving antenna is defined by:

$$A_e,\ r_x = \frac{\text{power delivered to the load of the Rx ant. } (P_L,\ r_x)}{\text{Incident power density } (P_{av},\ (\theta_{tx},\phi_{tx}))}$$

or, $P_L,\ r_x = A_e,\ r_x(P_{av},\ (\theta_{tx},\phi_{tx}))$

$$= [\ D_g(\theta_{tx},\phi_{tx})(e_t,\ t_x)]\ [D_g(\theta_{tx},\phi_{tx})\ )(P_{in,}\ t_x)\ (e_t,\ t_x)$$

or, $P_L,\ r_x = e_t,\ t_x\ \ e_t,\ t_x\ \left(\frac{\lambda}{4\pi R}\right)^2 D_g(\theta_{tx},\phi_{tx})(P_{in,}\ t_x)$

The above eq^n is known as fris transmission eq^n.

When,

$$e_t,\ t_x = e_t,\ r_x = 1$$

And both the receiving and transmitting are attig aligned in the direction of max^m rad^n, then.

$$P_L,\ \lambda,\ _x = \left(\frac{\lambda}{4\pi R}\right)^2 D_o,\ t_r,\ \ D_o,\ r_x\ \ P_{in},\ t_x$$

Here, the factor,

$\left(\frac{\lambda}{4\pi R}\right)^2$ is known as free space loss factor.

<u>Knife edge diffraction</u>:-

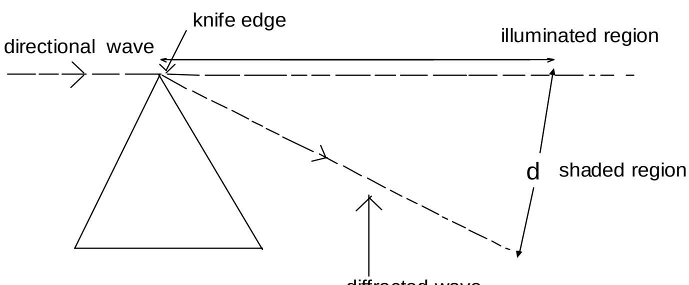


Fig. Knife edge difraction

The bending EM waves around the abstrades is known as diffraction and when edge then it is called knife as diffraction. Diffracted have is one that follows a path which can not be interpreted as either reflection or refraction. The following different parameter are useful to calculate, the relative power density in the shaded region .

Let,

P be the point of interest,

d= distance into the shadowed region.

r= distance from the knife edge into the projection of p on the boundary a separating the illuminated and shadowed regions.

A relative average power density (P_{av}, rel) usually estimated in the shadewed region influenced by the diffracted wave and is defined as:-

$$P_{av}, rel = \frac{\text{Avg. power density at a pt. of interest In the presence of knife edge}}{\text{Avg. power density at the same pt. in the absence of knife edge.}}$$

$$= \frac{\text{Avg. power density at a pt. of interest due to diffracted wave}}{\text{Avg. power density at the same pt. due to direct /reflected wave}}$$

$$P_{av}, rel = \frac{rd}{4\pi^2 d^2}$$

Where, it is assumed that,

$$\sqrt{\frac{2}{rd}} > 3, otherwise$$

$$P_{av}, rel = \left[\left\{\frac{1}{2} - \left(\sqrt{\frac{2}{r\lambda}}d\right)^2 + \frac{1}{2} - s\left(\sqrt{\frac{2}{r\lambda}}d\right)^2\right\}\right]$$

Where,

$$\left.\begin{aligned} C(x) &= \int_0^x Cos\frac{\pi u^2}{2}du \\ S(x) &= \int_0^x Sin\frac{\pi u^2}{2}du \end{aligned}\right\} \text{ are called fresnail integrals of cosine \& sine}$$

A table is provided for such integrals,

Thus, diffracti in which rel. avg power density is evaluated using fresnel integrals is called fresnel knife edge diffraction.

Chapter:- 4

Propagation in the radio frequency:-

Radio waves- The range of frequency between 104 HZ and 10^{10} Hz is generally referred to as radio frequency. The types of waves that radio frequency (RF) can take during propagation is depected in the following figure.

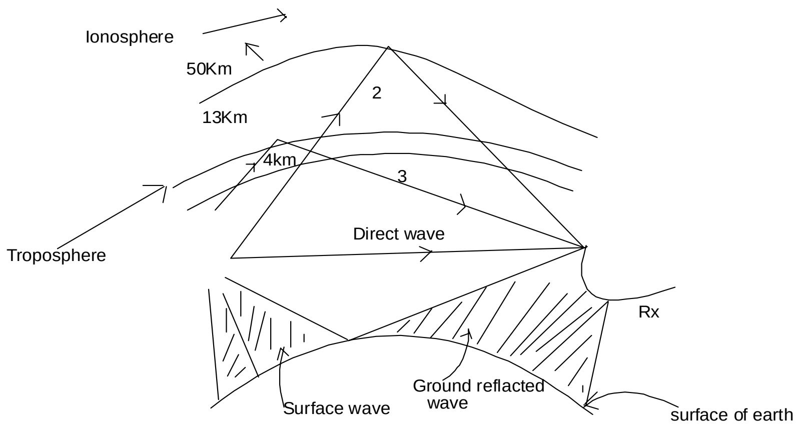


Fig.I Propagation of wave through different layer

(i) troposphere wave

(ii) – ionospheric wave/skywave

Here, direct wavw + ground reflected wave = specific wave.

Specific wave + surface wave = Ground wave.

Propagation through ionospheric reflection (sky wave):-

<u>Ionos sphere:-</u> The ionosphere is one of the layers of the earth's atmospheres which extends from about 50km above the earth to several of thousands kilometers. It is composed of no of ionospheric layers called D, E, and F. according to the free electron density present in that layer.

The free electron density is defined as the no of free electrons present in that layer per cm^3. since the electron density changes with time of day. Season, solar actimitres ectcs, the average electron density is obtained from a long observation . It is to be noted that E and F layer are the permanent layers whereas D layer exist only in day time. The lower region of the D layer. Sometimes exhibits the high

electron density and a separate layer known as a C layer has been added to this region.

In F layer, two separate layers are found during day time which are known as F_1 and F_2 layer.

The ionospheric reflection and the existence of the ionsphereic layers. Having different electron densities is explained as bnelow. The solar radiation couses the gas molecules present in the atmosphere to ionized with the liberation of the free electrons. At high altitude because the atmospheric is rare i.e only fewer gas molecules are available, very small no of free electrons are librated from the ionosation. Though the solar radiationis very strong. As the altitude decreases, the atmosphere gradually becomes denser but the solar radiation becomes weaker resulting in only small no of free electrons. A minimum electron density occur at a height of 50km from the earth surface. Between this two extremes. A considerable density of the gas molecules as well as the moderate intensity of the solar radiation exists approximately at height of about 300km above the earth surface resulting in highest no of free electrons i.e electron density.

Ionospheric reflection:-

As we know , in optics the refractive index ‘ η ‘ is defined by

$$\eta = \frac{\text{sine of angle of incident}}{\text{Sine of angle of refraction}}$$

$$= \frac{\sin\theta_i}{\sin\theta_r}$$

In wave theorem. The optical theories are equally valid and the refractive index in times of free electron density is given by:-

$$\eta = \sqrt{1-\frac{81N}{f^2}} \quad \text{\_\_\_\_\_\_\_\_\_\_\_\_(i)}$$

Where, N = free electron density present in the medium of which refractive index has been evaluated (in/m^3).

f = operating frequency (in Hz)

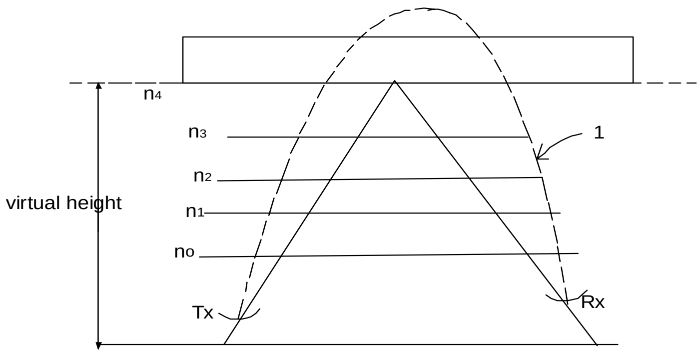


Fig. sketch showing the wave get totally reflected.

The radio wave passes through the layer with large electron density as it propagates up ward the ionosphere as seen from equation (i) the refractive indices n of the layers gradually decrease with the increase in ‘N’. This simply mean that the wave propagates from denser to the rarer ionospheric layers. This results in a larger angle reflection directly the refractive wave far away from the normal at the point of incidence. When the angle of refraction becomes 90^0 , the wave will be trailing horizontally. From fig (2) we can write, $n_0\sin\theta_i = n_1\sin\theta_1 = n_2\sin\theta_2 = n_2\sin\theta_3 = n_4\sin\theta_4$ and since.

$n_o = 1$, $\theta_4 = 90^0$

thus, $n_0\sin\theta_i = n_4\sin\theta_4$

or, $\sin\theta_i = n_4\sin 90^0 = n_4 = \sqrt{1-\frac{81N}{f2}}$

where, n_4 = free electron density present in the layer 4.

In general, $\sin\theta_I = n = \sqrt{1-\frac{81N}{f2}}$

From which, we get,

$$N = \frac{f2Cos^2\theta_i}{81} \quad - \quad --- \quad \text{(ii)}$$

So, if the electron density of the layer is sufficiently high to satisfy eq^n (ii), then the wave incident at the fixed angle θ_i with a certain frequency f_1, will be returned to the earth from that layer. If the maximum electron density in a layer is less them, that required by the equation(i), then the wave will penetrate the layer. However, it is noted that the same wave may be reflected back from the higher layer with electron density n sufficiently to satisfy the eqn (2).

From eq^n(ii), it can be set that when the angel of incident θi is zero (vertical or normal incidents). Then the maximum electron density Nmax will be required to make the wave reflected back to the earth. The frequency corresponding to Nmax

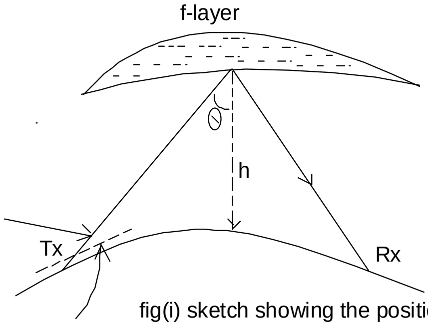
will be the highest frequency and is called the critical frequency for and is given by,

$$N_{max} = \frac{F(r)^2 Cos^2 0}{81}$$

$$F_{cr} = \sqrt{81 N_{max}} = 9\sqrt{N_{max}}$$

Thus we can say that each layer has it's own critical frequency. For the virtual height for a layer as shown in fig is that height from which a wave sent up at an angle appears to be reflected.

Maximum usable frequency:-(MUF)



fig(i) sketch showing the position of Tx

Fig:- (i) sketch showing the position of the Rx and height 'h' of the ionosphere.

From eqn (ii) and (iii)

$$f = \frac{\sqrt{81N}}{Cos\theta_i}$$

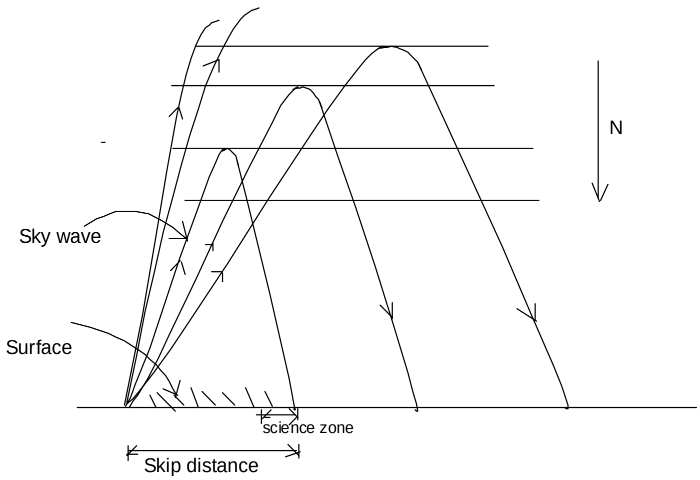
$$\therefore f = f_{er} / Cos\theta_i$$

This relation is called secant law.

It shows that the angle of incident is gradually increase the operating frequency f that the same layer would reflect back to the earth can also be increased correspondingly. . In practice, the angel of incident the transmitter and receiver and height of layer under considerations (usually f layer). As shown in fig (i). once the location for the TR and the receiver, and the layer of interest are decided, the operating frequency to the link also become fixed. This particular frequency is called maximum usable frequency (muf) for these two point. The term maximum is use in this because the frequencies, higher then MUF would not be reflected form the layer of under consideration. From eq^n (ii) for a constant angle of incident as the frequency is increased, the layer must also posses higher

electron density to reflect their higher frequency back to earth. Since the layer under consideration already fixed, it just fail to do so, therefore the wave passes through the layer and may scope into the space only the required electron density is available in the path.

Again referring to eq^n (ii) for constant operating frequency, the angle of incidence when decrees. The required electron density to make the wave . Reflect back to the earth is increase if the required electron density is present in the ionosphere the wave gets reflected, it escape into the space otherwise. Again when angle of incident is increased, the required electron density is decreased and the wave get reflected back to the earth but not necessary to the target because the layer or layers with lower electron density are always present in he iososhpere . This case is depted (shown) in figure below:-



The shortest distance measure along the surface of the earth between the transmitter and the point at which the sky wave of fixed frequency meets the ground is called skip distance. Hence MUF can be defined as , that frequency which makes the distance between the two point of communication as the skip distance. The reception is possible within the certain distance from the transmitter due to the surface wave . The area length uncovered with both the surface and sky wave is called the silence wave where no reception is possible. The MUF attains a maximum value when the wave is tangentially transmitted. i.e the angle ψ shown as grazing angle become minimum (i.e $\psi = 0$). In this case, f = 3.6fcr.

$$f = 3.6f_{cr} \qquad \text{(v)}$$

In real life it is observed that MUF varies about the monthly average of upt to 15%. So the operating frequency in practice is choose about 15% less them MUF. And with the field strength falling of about 10 to 20 dB. The frequency is increased from operating to the MUF.

Lowest usable frequency(LUF):-

Although the frequency higher then the MUF can not be use, the frequency lower then MUF can still be use to link the same two points using suitable angle of incidence, and the ionospheric layer. But is results in lower field strength because ionosphere absorption increases as the frequency decreases. This decrease in the field strength can be computed only to a certain extent by increasing the transmitter power. Therefore, there a unit in using the lower frequencies as well . The lowest frequency that can be use to produce the satisfactory reception for a given distance and the transmitting power is called the lowest usable frequency.

Very high frequency (VHF) propagation (space wave):-

The VKF (30 to 300 MDZ) is used for F.M sound broadcasting, line of sight communication and TV broadcasting . when the operating frequency F lies within the range of HF (3 to 30 MHZ) the waves usually reflect from the ions phere and comeback to the earth. This leads the ionspheric wave to be the dominant wave. Now if the operating frequency falls in the range of VHF then the space wave will be the main wave because ionspheric wave component of VHF pulses through the ionosphere into the space and the surface wave component is strongly observed by the ground . Besides the VHF is almost completely obstructed by the abstractacs such as building and trees.

The electric field strength at the receiving it is given by

$$E = \frac{240\pi I_o H_{tx} h_{rx}}{\lambda d^2} Vm^{-1} \quad \text{provided } d >> h_{tx} + H_{rx}$$

here,

I_o = max^m current in the T_x antenna.

h_{tx} = effective height of the T_x antenna.

h_{rx} = effective height of the T_R antenna.

d = distance between the antenna. (miles)

Tropospheric wave (tropospheric scatter propagation):-

When the operating frequency falls in the range of UHF (300 mHZ ot 3 GHz). The main component of the wave will be the trospheric wave or scatter wave. Usually the electron density ‘N’ increase as the height of atmospheric layer above the earth surface increase resulting in lower refractive index of the layer. Because

this gradual reduction in the refractive index is not neigh to bend the wave of there frequency, the wave therefore passed through all the layer scaping into the space. But the patches of highly ionized regions called the blobs are found in the trosphere. This makes the change in the refractive index very rapid which is enough to bend UHF back to the earth surface resulting in a useful wave of communication. The reason behind this type of communication is not understood completely but also it is believed that the wave undergoes scattering similar to the scattering of search light beam. This phenomenon is a permanent phenomen and links of 300 to 500 km are typically used. The power at the receiving point is given by the.

$$P_{rx} = 1.8 \times 10^{32}\ P_{tx}\ \beta^{-5/3}\ G_{tx}\ G_{rx}\ C^2_n\ \frac{\theta^2_{HP}, t_x}{d^{(17/3)}} \text{ watts}$$

where, P_{rx} = power received by the receiving antenna.

P_{tx} = Power transmitted by the T_x antenna.

β = Phase constant

G_{tx} = gain of the T_x antenna.

G_{rx} = gain of the R_x antenna.

$C_n = 5 \times 10^{-7.5}$ = a constant

θ_{HP} = half power of T_x antenna.

d = distance between the antenna (d << radius of easth)

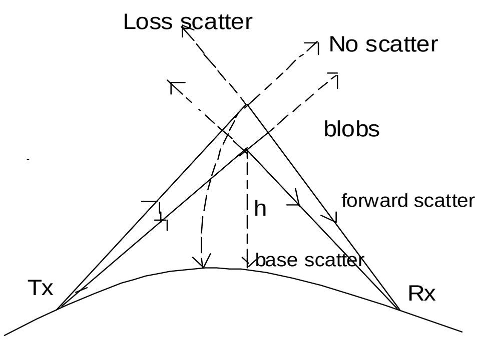


Fig. (i) Tropospheric progagation

The propagation loss estimated as the ratio of power received vcia to that obtained with friss equation formula and is found to be in the range of $1/10^5$ to $1/10^7$ or (-60 to -50 dB). Because of very high loss the transmission of very high power are often required for reliable communication. This results the cost to be approximately four times the co-=axial transmission and 12 times the microwave links. This type of propagation is extremely useful over rough or enable terrain (area) inspide of it's high cost.

Microwave propagation:-

The electron density of atmospheric layer increases with it's height above the earth surface. Thus the refractive indexes of the layer gradually decrease. This gradual reduction in the refractive index is not enough to bend the microwave back to the earth surface. Hence the wave with this frequency pass to the atmosphere and scope into the space. But under certain atmospheric condition a layer of warm air may be trapped above cooler air ; often over the surface of water the result is that the refractive index will decrease far more rapidly as the height inch area. This happens near the ground often within 30m from it.

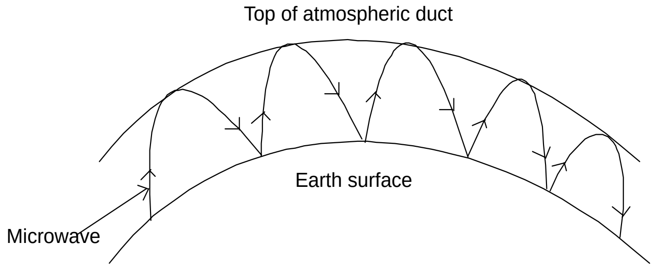


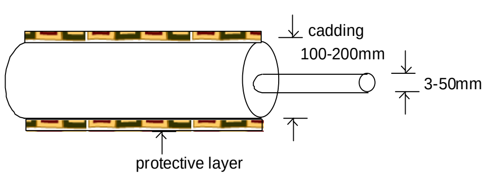
Fig. microwave propagation through superfractino or ducting

The rapid reduction in the refractive index will do to microwaves what the slower reduction of their quantities in an ionspheric layer does to Hf wave.

A complete bending down therefore takes place as shown in fig. above . Microwaves are thus continuously refracted in the duct, and reflected by the ground , so that they are propagation around the curvature of earth for distances, which sometimes, exceed 1000km. This phenomens is also called the super fraction or ducting. The main requirement for the formation of atmospheric ducts is the so- called temperature inversion. This is an increase of air temperature with height, instead of usual decrease in temperature with 6.5^0 c/Km in the standard atmosphere.

Chapter:- 5

Optical fibre:-



Basic of light propagation:-

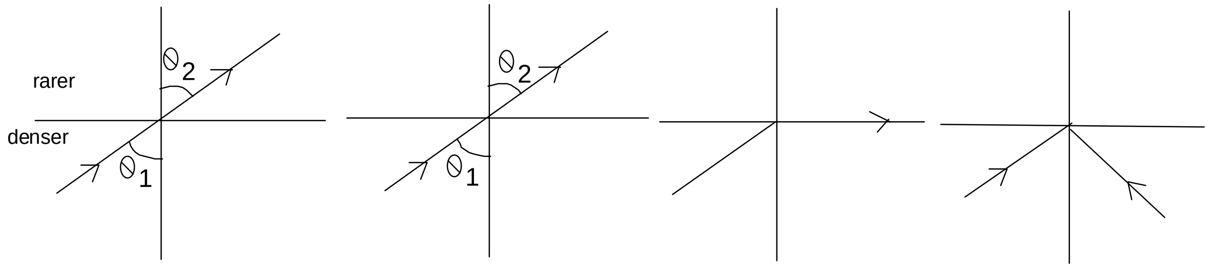
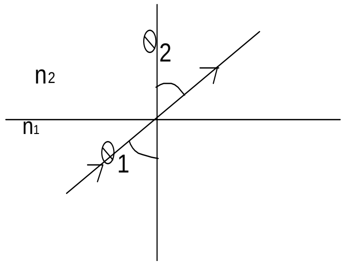


Fig. TIR



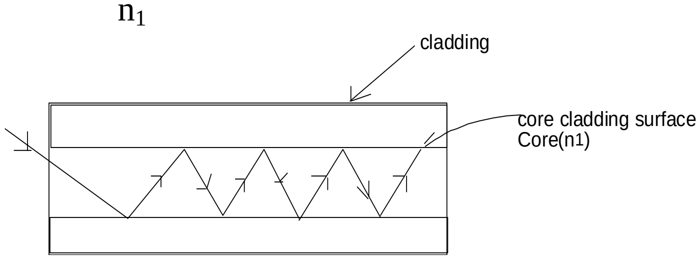
$(n_1 > n_2)$

$n_1 \mathrm{Sin}\phi_1 = n_2 \mathrm{Sin}\phi_2$

Now, when $\phi_1 = \phi_2$

$n_1 \mathrm{Sin}\phi_c = n_2 \mathrm{Sin}90^0$

$\mathrm{Sin}\phi_c = \dfrac{n_2}{n_1}$ Snell's law

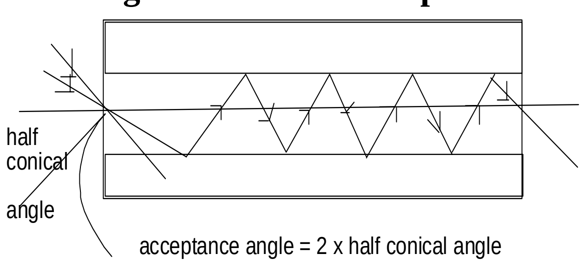


rays → meridinal rays
rays → skey rays

meridinal rays(through axially)

skey rays (in helical form)

Acceptance angle and Numerical aperture:-



Numerical Aperture (NA):-

- Light collecting ability
- Establishes relation between on medium and refractive indices.
- From optics theory $n_o \sin\theta_o = n_1 \sin\theta_1$

  h_o = refractive index of medium

  θ_o = acceptance angle (half conical angle)

  h_1 = refractive index of Core.

  θ_1= angle of refractive (air –core interface)

$n_o \sin\theta_o = n_1 \sin(90^0 - \phi_1)$

$n_o \sin\theta_o = n_1 \cos\phi_1$

$= n_1 \sqrt{1 - Sin^2\phi_c}$

$n_o \sin\theta_o = n_1 \sqrt{1 - Sin^2\phi_c}$

for air n_o= 1 and for critical condition.

$\phi_1 = \phi_c$

$\sin\theta_o = n_1 \sqrt{1 - Sin^2\phi_c}$

$= n_1 \sqrt{1 - Sin^2\phi_c}$

$\sin\theta_o = \sqrt{n_1^2 - n_2^2}$

$\therefore NA = \sin\theta_o = \sqrt{n_1^2 - n_2^2}$

-relative refractive index between core and cladding is

$$\Delta = \frac{n_1^2 - n_2^2}{2n_1^2}$$

$\Delta << 1$

$$\Delta \approx \frac{n_1 - n_2}{n_1}$$

$\therefore NA = \sin\theta_o = n_1 \sqrt{2\Delta}$

$\therefore NA = n_1 \sqrt{2\Delta}$

Type of optical fibre (on the basis of refractive indices profile)

(i) Step index fibre.

(ii) Graded index fibre.

(i) Step index fibre:-

$$n(r) = \begin{cases} n_1 & r < a \text{ (core)} \\ n_2 & r \geq a \end{cases}$$

(ii) Graded index fibre:-

$$n(r) = \begin{cases} n_1\sqrt{1-2\Delta(r/a)^2} & r< a \quad \text{(core)} \\ n_2 & r\geq a \quad \text{(cladding)} \end{cases}$$

α = 1 Triagular shape (profile)

α = 2 parabolic shape

$\alpha = \infty$ step profile

Page 97:- preparation of optical fibre

Sidy (vapour) + 2 H_2O (Gas) --- SiO_2 (solid) + HCL (gas)

Sidy (vapour) + O_2 (Gas) --- GeO_2 (Solid) + $2CL_2$ (gas)

Number of modes in fibre:-

TE_{nr} = Transverse electric ($E_z = 0$) mode

TM_{nr} = " " " ($H_z = 0$) mode

HE_{mr} = Hybrid (E_z #0, H_z#0) mode (H>E-Field)

EH_{nr} = Hybrid (E_z #0, H_z# 0) mode (E>H- Field)

Where, n = n^{th} order of Bessel function

r = order of root of n^{th} order Bessel function.

V- number:-

⇒ It represent the number of modes.

$$V = \frac{2\pi}{\lambda} a(NA)$$

For this V,
The number of modes in step index (NS) is,

$$M_S \approx V^2/2$$

And in Graded index (M_G)

$$M_G \approx \left(\frac{\alpha}{\alpha+z}\right)\frac{V^2}{2}$$

<u>The desingnation of mode step index fibre:-</u>

$M_s = 1$ ⇒ Single mode step index fibre.

$M_s \geq 2$ ⇒ Multimode step index fibre.

$M_G = 1$ ⇒ Single mode Graded index fibre.

$M_G \geq 2$ ⇒ Multimode Graded index fibre.

Light sources and detectors:-

- optical sources (EOC)
- optical detectors (OEC)
- e.g of optical sources :- (i) led

(ii) laser

e.g of optical detector:- (i) photodiode

(ii) p-i-n diode

(iii) Avalanche diode.

Properties of optical sources:-

(i) compatible size
(ii) tracking of input signal
(iii) Emit light at wavelength.
(iv) Direct modulation can be implemented. (from few range of frequency to mHz).
(v) Couple sufficient optical power.
(vi) Narrow spectal Band width.
(vii) Stable optical o/p.
(viii) Cheap and highly variable.

Chapter- 2

Aperture antenna:-

An antenna having a aperture (opening) with certain geometrical shapes is referred to as an aperture antenna. It may take the form of wave guide or a horn. A horn is hollow pipe of different crossectional which has been tapered to a large opening . The opening may be square, rectangular, ellipse, circular etc. The aperture antenna operate at microwave frequency (greater than or equal to 1 GHz) and therefore they are also referred as microwave antenna. Fig (i) below shown the different form of aperture antenna.

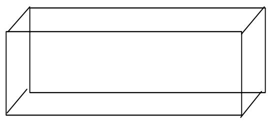


fig. a. rectangular aperture

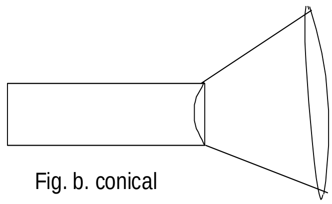


Fig. b. conical

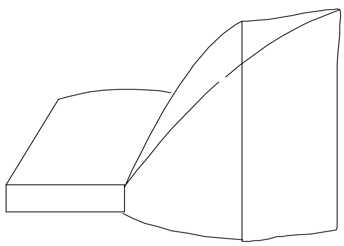


Fig.C Rectangular

Need of aperture antenna:-

Aperture antennas are most widely used in real life because:-

(i) they have larger gain.

(ii) They are easily flush mounted to the surface of space craft or cir craft without disturbing the aerodynamic of the craft.

(iii) They are convenient to be covered with a dielectric to protect them from unforourable environment continuous.

(iv) Act as a very suitable feeding element for other antenna such as reflect or antennas.

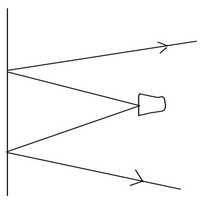
When the aperture of the horn takes the form of rectangle, the resulting antenna is called the rectangular horn antenna of the opening is flared in the direction of E/H field, then the horn is said to be E/H horn and when opening is flased in the direction of e and H field , the horn is called pyramidal horn.

Reflection antenna:-

The reflector antenna may take different forms according to the shape of the reflector use. Because reflector antennas have high gain, they are usually used in radar application, satellite communication and other microwave communication. Among the different types the front fed parabolic reflector are most widely use large aperture ground based antennas.

The world largest this type of reflector which is located in most Germany is 120m in diameter and in USA; it is 64m in diameter.

The horn or wave guide operating single mode are mostly use as feeder for the reflector antennas. Since a feeder is placed at the focal point of the parabola the configuration is usually known as front fed. The waves that eminates (from the reflector in parallel fashion is termed as collimator). The various form of reflector antennas are as follows:-



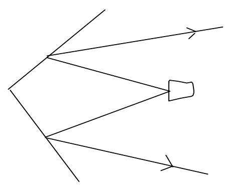
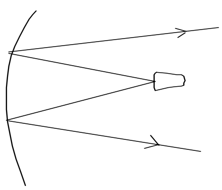
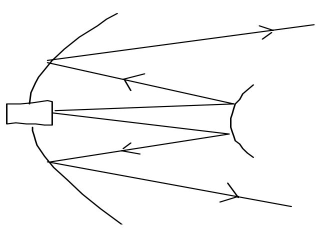


Fig (a) plane reflector b. corner reflector



© fronifed parabolic reflector



(d) Cass grain fed parabolic refractor.

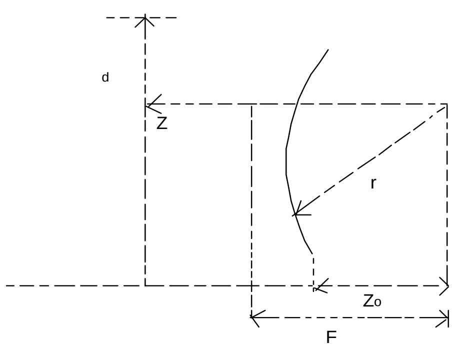


fig. (i) the shape geometry of a front fed parabolic reflector

One of the frequently used parameter of the reflector is the subtended. angle and it is given by,

$$\theta_0 = Tan^{-1}\left(\frac{d/2}{z_o}\right)$$

$$= \tan^{-1}\left[\frac{1/2(f/d)}{(f/d)^2 - \frac{1}{16}}\right]$$

The directivity D_o is given by

$$D_o = \left(\frac{\pi d}{\lambda}\right)^2 \left\{Cos^2(\theta_o/2)\left|\int_0^{\theta_o} \sqrt{Gf(\theta')}\tan(\theta/2)d\theta'\right|^2\right\}$$

$$Gf(\theta') = \begin{cases} \{G_o{}^n Cosn(\theta')\} & 0 \le \theta' \le \pi/2 \\ O & \pi/2 < \theta' \le \pi \end{cases}$$

or, $D_o = \left(\frac{\pi d}{\lambda}\right)^2 \in_{ap}$

where, $\in_{ap}$ = aperture efficiency.

Numerical :-

Question:- Calculate the maximum usable frequency if electron density of ionosphere is $1.20*10^{12}$ electron /m^3 and the angel of incident is 40^0.

Sol^n:-

Given, $N = 1.20*10^{12}$ electron/m^3

$\theta_i = 40^0$

MUF = ?

We know that the freq. corresponding to MUF is, F = Fer sec

F = fer sec θ_I = $\sqrt{81N}Sec\theta_i$

$= \sqrt{81*1.2*10^2}Sec40^0$

= ……………..

Question:- 2

A step index glass fibre has higher core index and lower cladding index of 1.05 and 1.45 . Find out the following parameters of such a fibre:-

(i) Critical angle of the fibre.

(ii) Corresponding acceptance angle of the fibre.

(iii) Numerical aperture.

(iv) % of the light collected in the fibre,

Sol^n- (i) n_1 = - 1.50

$n_2 = 1.45$

Critical angle (θ_c) is given by,

$\theta_c = \text{Sin}^{-1}(n_2/n_1)$

$= \text{Sin}^{-1}(1.45/1.50)$

(ii) Since the acceptance angle of the fibre is form air to core.

$n_a \text{Sin}\theta_c = n_1 \text{Sin}\theta_c$

But, (n_a =1 for air)

$\text{Sin}\theta_c = \text{Sin}^{-1}(1.5\text{Sin}\theta_c)$

= …………..

$$NA = \sqrt{n_1^1 - n_1^2}$$

$$= \sqrt{15^2 - 1.45^2}$$

$$= \sqrt{2.25}$$

(iii) The percentage of the light collected by fibre is given by:-

$$(NA)^2 \text{ x } 120\%$$

(iv) A silica optical fibre with a core diameter larger enough to be considered by ray theory analysis has core refractive index has 1.5 and cladding refractive index of 1.4 # determine.

a. The critical angle of the core cladding interface.
b. NA for the fibre.
c. The acceptance angle in air for the fibre.

Notes:-

(i) Calculation of MUF:-

Case- I

Thin layer (or Flat earth)

$$f_{MUF} = f_c \sqrt{1 + \left(\frac{D}{2h}\right)^2}$$

where,

f_c= critical freq.

d = distance between T_x & R_x.

h = height or virtual height.

Case- II

Thin layer (or curved earth)

$$f_{MUF} = \frac{fc\sqrt{\frac{D^2}{4}+\left(h+\frac{D^2}{8R}\right)^2}}{\left(h+\frac{D^2}{8R}\right)}$$

where, R = radius of the earth.

(2) The skip distance Dskip is given by:-

$$Dskip = 2h\sqrt{\left(\frac{f_{MUF}}{fc}\right)^2-1}$$

(3) The line of sight (Los) distance or range of space wave propagation is given by:-

$$d = 3.57[\sqrt{h_t}+\sqrt{h_r}]\ km$$

where, ht & hr are heights of transmitting & receiving antennas in meter and d is in km.

Question:- A high frequency radio link has to be established between two points at a distance of 2500km. considering inospheric height to be 200km and critical frequency 5MHz. calculate. MUF.

D= 2500km = 2500 x 10^3m

h= 200km = 200 x 10^3m

fc= 5MHZ = 5 x 10^3HZ

$$f_{MUF} = fc\sqrt{1+\left(\frac{D}{2h}\right)^2}$$

$$= 5\times10^3\sqrt{1+\left(\frac{(2500\times10^3)}{2\times200\times10^3}\right)^2}$$

Question :- Find the MUF for a distance d= 1.3mm by fz layer (h= 325km) reflection with f2 layer electron density. N= 6 x 10^{11} mc^{-3}. Neglect earth curvature.

Question:- A $\lambda/2$ transmitting antenna radiates 10 kw of power at 100mHZ. if the height of transmitting and receiving antenna are 100m and 9m respectively, calculate the radio horizon range in km. also calculate he field strength at the receiving end.

Sol^n:-

The field strength at the receiving end is given by.

$$E_R = \frac{88\sqrt{p}\,h_t h_r}{\lambda d^2} v/m$$

Where, P = power Tx mitted = 10Kw

F = 100 mHz

$$\therefore \lambda = C/F = \frac{3\times10^8}{100\times10^6} = ...$$

ht = 100m

hr = 9m

d = ?

we know,

$$d = 3.57\times10^3[\sqrt{h_t}+\sqrt{h_r}]$$

$$= 3.57\times10^3\times13$$

$$= 46.41 \text{ km.}$$

Question:- If the maximum density in the ionosphere corresponds to a refrective index of 0.82 for a frequency of 10mHz . Determine, neglecting the effect of earth magnetic field, the ground range for which this frequency is maximum usable frequency. Assume that the reflection takes place at a height of 300km and the earth is a flat surface.

Sol^n:- $f_{MUF} = f_1\sqrt{1+\left(\frac{D}{2h}\right)^2}$

$$\text{or, } D = \sqrt{\left(\left(\frac{f_{MUF}}{fc}\right)^2-1\right)}2h$$

where,

$$fc = 9\sqrt{N}$$

$$\& \; n = r.1 = \sqrt{1-\frac{81N}{f^2}}$$

$\Rightarrow$ N = ?

Problems:-

(i) Transmission loss

(ii) Radio wave propagation

(iii) Multiplication of pattern.